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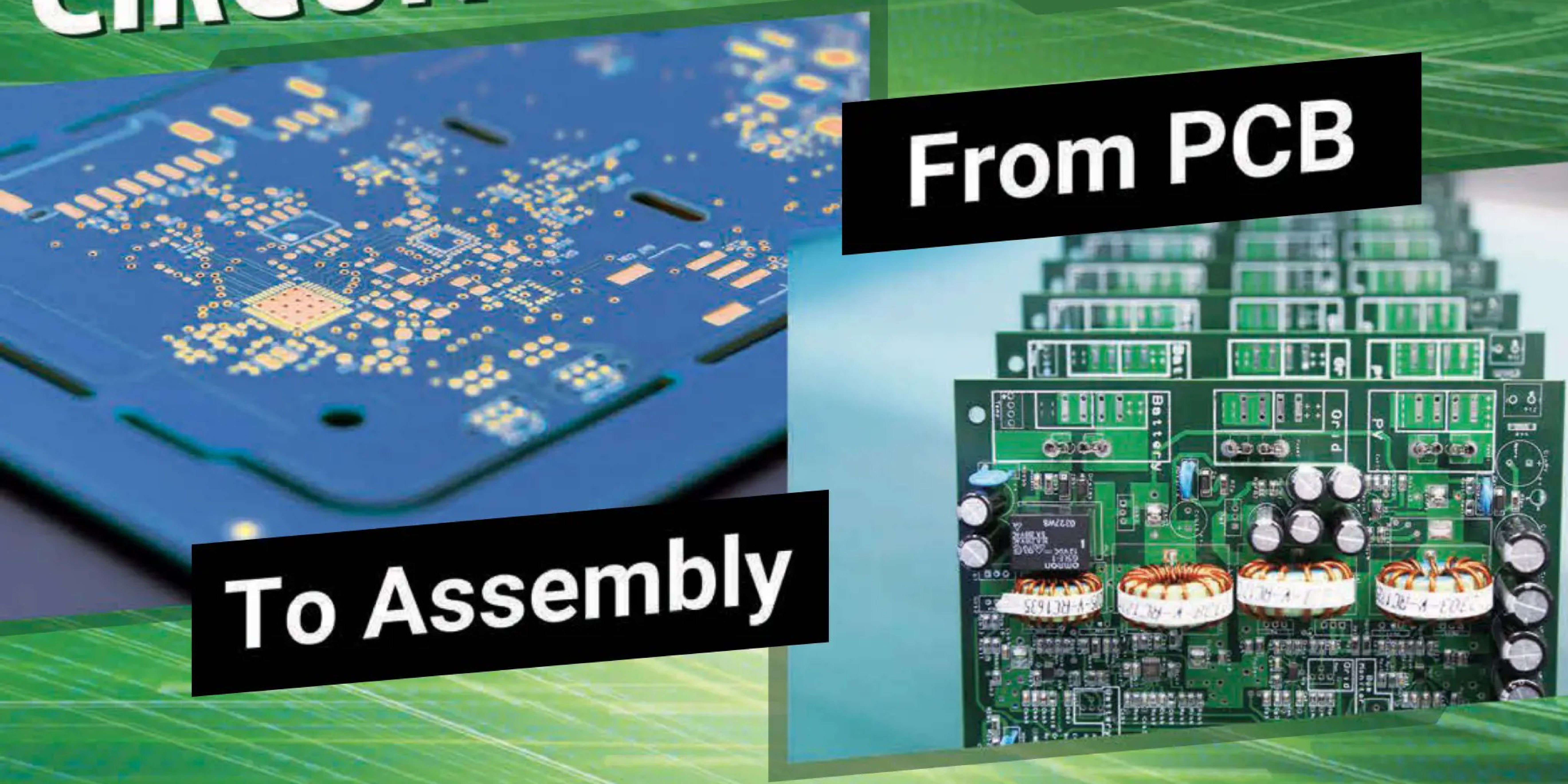

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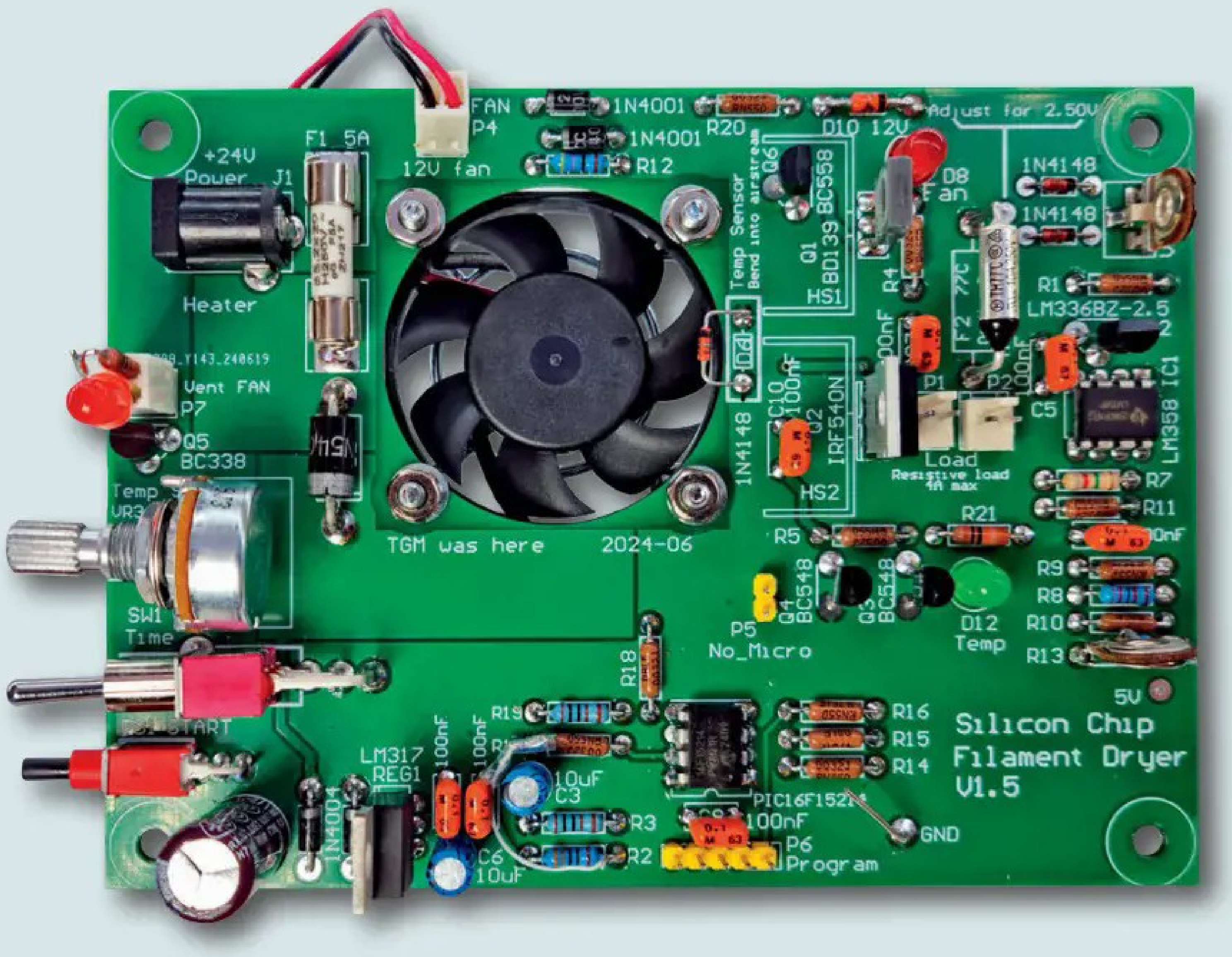
Contents



Projects and Circuits

Seaside Sound Simulator by John Clarke 6
Relax and enjoy the sound of the seaside from the comfort of your home. Forget the scorching heat in summer, the cold winds of winter or your swimsuit being full of sand! It's a fun-filled project, ideal for beginners and experienced constructors alike.

Compact HiFi Headphone Amplifier, part two by Nicholas Vinen 27
Our new Stereo Headphone Amplifier fits in a neat package and has two sets of inputs with individual volume controls. Having described its performance and how it works, we'll look at the circuit board design before getting into construction and testing.

Upgraded Electric & Bass Guitar Pickguards by Brandon Speedie 42
The control circuitry on electric basses has remained remarkably simple for over 70 years. These PCBs offer more advanced features and a cool aesthetic. They suit many popular models of electric bass, as well as the Fender Telecaster electric guitar.

3D Printer Filament Drying Chamber, part one by Phil Prosser 60
This enclosure can store up to four 1kg reels of 3D printer filament, keeping them dry and ready for use at any time. You don't even need to remove them – the filament can simply be fed to the printer through a small hole in its lid.

Series, Features and Columns

Techno Talk by Max The Magnificent 4
Banshees & Bananarama; join me on a journey through music technology history.

Audio Out by Jake Rothman 14
Three discrete class-A buffer module PCB designs.

Precision Electronics, part nine by Andrew Levido 20
This final instalment in the series brings together everything covered in the first eight parts with its discussion of overall System Design. We take what we have learned so far and see how to approach the design of a precision electronic system.

Max's Cool Beans by Max The Magnificent 34
Weird & Wonderful Arduino Projects, part 10: How good are your reactions?

The Fox Report by Barry Fox 41
Electronics retailers stand to benefit from the end of Windows 10 support in October.

Circuit Surgery by Ian Bell 52
Measuring the frequency response of a circuit using a PC sound card, part 1.

The Life of Nikola Tesla, part one by Dr. David Maddison 68
Nikola Tesla was a prolific inventor, engineer, futurist and essayist. He spoke eight languages and had a wide range of interests. Despite his 'mad scientist' vibe, his contributions to our modern industrial civilisation are significant.

Regulars and Services

Editorial 2
Are surface-mount devices less reliable than through-hole parts?

Reader services – Editorial and Advertising Departments 2

Subscribe to Practical Electronics and save money 3

Practical Electronics Back Issue DOWNLOADS – 2024 now available! 33

Electronics Teach-In Bundles 59

Practical Electronics PCB Service 76
PCBs for Practical Electronics projects

Classified ads and Advertiser index 79

Next month! – highlights of our next issue of *Practical Electronics* 80

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Editorial

Are surface-mount devices less reliable than through-hole?

This month, I became aware of the idea that some people think that electronic devices populated with surface-mount devices (SMDs) are somehow less reliable than those made with the older-style through-hole parts.

My first thought was to question why the packaging method of a device affects its reliability. For example, the good old TO-92 package BC546 is available in an SOT-23 SMD package, as the BC846. It has a lower dissipation rating of 310mW compared to 625mW, due to its smaller size, but the internals are identical. Other than failing to take that lower dissipation rating into account (which would be a design flaw, not a fault of the part), is there any reason to believe that a BC846 should not last as long as a BC546?

In fact, SMDs were developed in the 1960s primarily for military and aerospace applications specifically because they were **more reliable** (and also more compact, which obviously helps when you're trying to fit electronics in a rocket or aircraft). That's because lead flex means that a component body on the end of leads will tend to move around in response to vibration and shocks. That stresses both the leads and the solder joints, risking eventual failure.

That doesn't mean SMDs are always better; they are more sensitive to board flex, for example. But in general, they are thought to be more reliable, are cheaper to make, take up less space, offer better RF performance due to the smaller size and so on. Of course, being smaller means you need better eyesight to see and work with them (unless you're using a microscope), steadier hands and finer tools.

The myth of SMDs being unreliable likely persists because poorly-engineered budget devices use them – not because the parts themselves are inherently inferior. When cheap gear fails, people open it up and see SMDs inside, and assume a causal link.

When it comes to repairing devices, there are actually significant advantages to SMDs. Yes, you need more skill to work on them, but you can remove and reinstall them with access to just one side of the board. It's also possible to ruin multi-layer boards with plated through-holes when removing through-hole parts (and it's frustratingly difficult, too).

Ultimately, the biggest obstacle to repair these days is lack of documentation or replacement parts, not the difficulty of replacing the failed components once you've identified them.

I understand why people might not like to work with small, fiddly parts but they are not inherently any less reliable.

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Banshees & Bananarama

Techno Talk

Max the Magnificent

From ancient pots playing prehistoric hits to sirens wailing in the wee hours of the morning, join me on a journey through music technology history—because every groove deserves to be remembered.

***The Nightside* is a series of 12** books by British author Simon Green. This refers to a hauntingly surreal, neon-noir pocket dimension nestled within London, where the supernatural and science fiction super-technology coexist, and it's always three o'clock in the morning—the "hour of the wolf"—when the thin veil between reality and the supernatural hangs in the balance.

I was just re-reading the third book in the series, *Nightingale's Lament*. One line that caught my attention was when our hero, John Taylor, who is talking about some of the more nefarious nightclubs, says, "There are any number of ways in which singing can bring about horror and death: sirens, undines, banshees, Bananarama tribute bands…"

Now, there's a man who knows what he's talking about! For some reason, this set me to ruminating on the wide variety of technologies humans have invented to record and replay music over the years.

That's potty!

I vaguely remember reading a short story in a science fiction anthology many moons ago. As I recall, a team of archaeologists discovered a jar created on a potter's wheel in Mesopotamia around 3000 BCE. The outside of the jar was decorated from top to bottom with a continuous spiral groove composed of concentric incised lines.

They determined that these lines were created by the potter holding a sharp-pointed stick (or stylus) lightly against the surface of the clay as the pot rotated on the wheel. When they looked closely, they saw that the groove contained modulations, similar to those found on modern vinyl records.

They created an apparatus to read the modulations and recreate the sounds, at which point they heard people talking and the potter singing.

I don't know about you, but I find the idea of listening to conversations from thousands of years ago to be strangely compelling, even though it's no different in concept to listening to recordings from the 1920s, for example.

Although there is (as yet) no credible evidence of ancient ceramics preserving unintentionally recorded sound, I prefer to think that there remains a tantalising possibility.

I told them we've already got one!

Based on our high school history lessons, most of us are at least vaguely aware that the American Thomas Alva Edison created the first phonograph—a machine capable of recording and playing back sound—in 1877.

Less commonly known is that the French inventor Édouard-Léon Scott de Martinville invented a device called the phonautograph—a machine that could record sound, although it couldn't play it back—in 1857, 20 years before Edison.

You may feel that the ability to record sound but not replay it is as helpful as a write-only memory, or WOM (**https://w.wiki/vNB**). However, the phonautograph recorded sound as "squiggles on paper", thereby providing the first method for humans to visualise sound.

Even better, and in keeping with our musings on ancient pottery, in 2008, scientists managed to convert a phonautograph recording captured in 1860 back into sound (**https://w.wiki/Ez43**).

Cylinders vs discs

Edison's original phonograph used tinfoil on a rotating cylinder, with the groove running in a helical spiral around the cylinder's surface. In the late 1880s, German-American inventor Emile Berliner introduced the gramophone, which used flat discs with a spiral groove that started near the outer edge and moved inward.

As we know, Berliner's disc format eventually dominated the market, but it certainly wasn't a case of being 'done and dusted'. During the first half of the 20th century, records were made from a natural resin called shellac. This was heavy and brittle, allowing speeds to range from 60 to 120 revolutions per minute (RPM), with diameters ranging from 7 to 12 inches (~18 to 30cm).

By the 1950s, the market had transitioned from shellac to vinyl (polyvinyl chloride or PVC resin), with the primary formats settling on 12-inch (~30cm), 33⅓ RPM LP (long play) records for classical music and full albums. 7-inch (~18cm), 45 RPM discs were used for single songs, including in jukeboxes.

Records prior to the 1950s supported only a single audio channel. During the 1950s, stereophonic sound—with separate left and right channels—became the mainstream format. The ability to encode two channels in a single groove is a testament to the ingenuity of mid-twentieth-century engineers.

Stereo encoding on vinyl uses a mechanical scheme similar to 'mid-side' coding. The groove is V-shaped, typically with a 90° included angle, and its two walls are cut at ±45° to the vertical. Sound for the left channel is encoded as movement of the stylus parallel to the left wall; sound for the right channel is encoded as movement parallel to the right wall.

When combined, lateral (side-to-side) motion corresponds to L + R (the mono sum), while vertical motion corresponds to L – R (the stereo difference). A mono cartridge responds only to lateral movement, automatically ignoring the stereo difference, which is why a mono record player can play stereo records without trouble.

This approach mirrors the transition from monochrome to colour television; the brightness (luminance) signal was transmitted exactly as before, while the colour (chrominance) signal was added in a way that monochrome sets would ignore. As with stereo records, the new format was compatible with existing equipment.

Backwards compatibility of this sort is rare today, often dismissed as unnecessary—or worse, as an obstacle to planned obsolescence.

Wires vs tapes

Another interesting technology for recording and reproducing sound is magnetic tape. The first reel-to-reel machines originated in the late 1920s and early 1930s, but these were far too expensive for general consumer use.

Automobiles provided a big driver (no pun intended) for magnetic tape. Car ownership started to boom in the 1950s. People began to spend more time driving around and wanted to listen to music while doing so. Radio was nice enough, but you had to listen to whatever was served, and people wanted to listen to what they liked.

In 1962, the American businessman and engineer "Madman" Muntz created a 4-track cartridge that involved an endless loop of tape. Two years later, William Powell Lear led a consortium of companies to develop the 8-track cartridge and player, which took both automotive and home users by storm.

Later, in the 1970s, compact cassettes started to gain ground, and this format really took off with the introduction of the first Sony Walkman in 1979.

We could waffle on about tape technology for yonks, but we have other fish to fry. For example, I feel it would be remiss of me to fail to mention magnetic wire recorders, the first of which was invented by Danish engineer Valdemar Poulsen in the late 1890s.

I first heard about these a few years ago, when Nick Gent emailed me to tell me an amazing story about some wire recordings he'd found that were made by his grandfather (**https://pemag.au/link/ac81**).

I love my H2O bone conduction headphones.

Files vs feeds

There's so much we've left uncovered, like the ascent of CDs in the mid-to-late 1980s, for example. In turn, these were supplanted by MP3 players and digital downloads in the late 1990s and early 2000s. Today, most people prefer to stream music rather than download it, with streaming currently accounting for ~85-90% of all music revenue.

*[Editor's note – like vinyl records, CDs are making something of a comeback these days, although they never really left – see **https://pemag.au/link/ac82**]*

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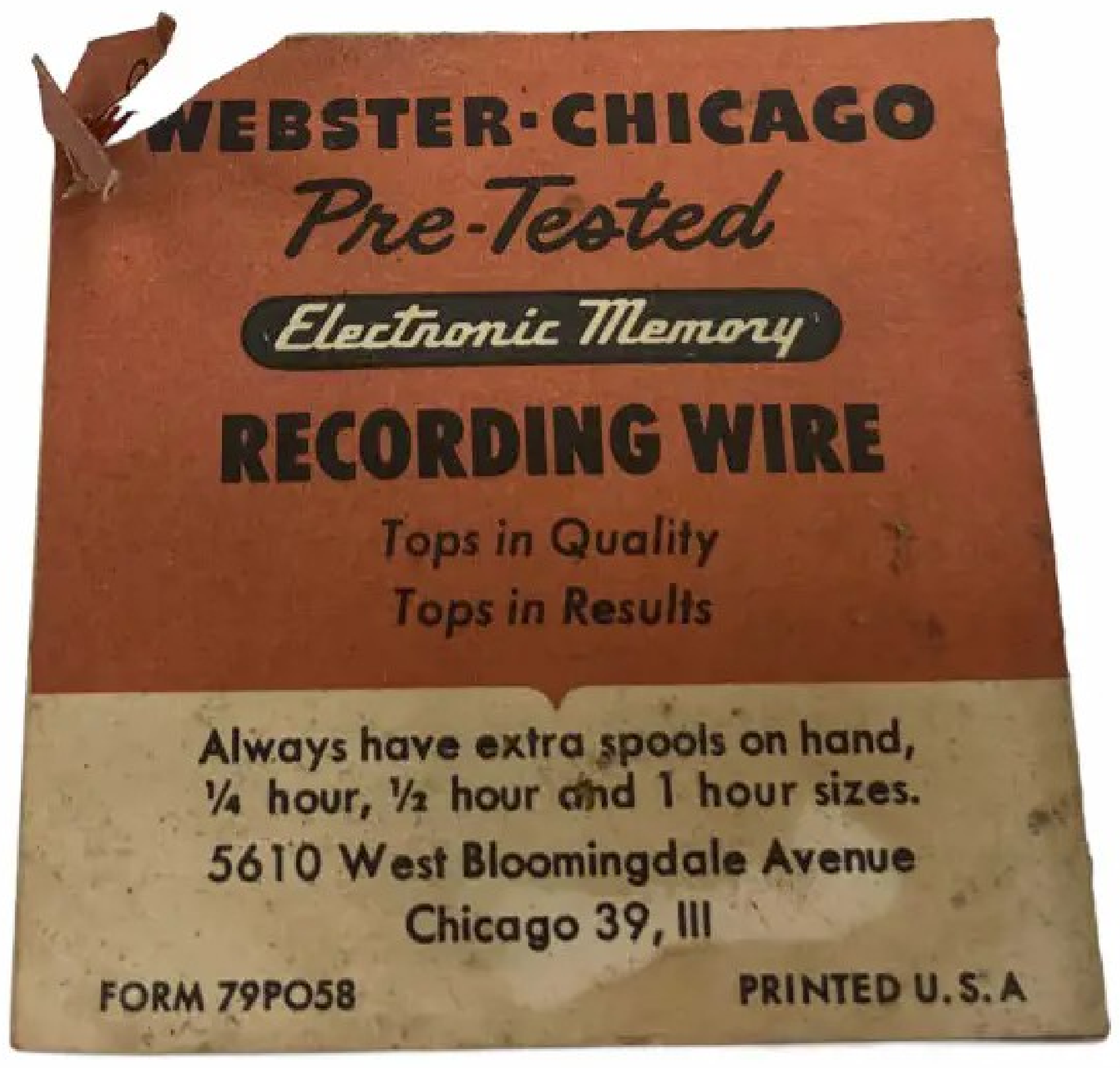
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As an aside, most people assume that streaming music involves the MP3 format because that's all they've heard of; however, most streaming services utilise more modern coding and compression formats, such as AAC (advanced audio coding), Ogg Vorbis (an open-source codec from **Xiph.Org**), Opus (also open), FLAC (free lossless audio codec, also from **Xiph.Org**) and ALAC (Apple lossless audio codec).

Bad to the bone!

As one final tidbit of trivia: most people think of headphones or earbuds when they want to listen to audio on the move. However, I prefer to maintain situational awareness in the event of something unfortunate happening, such as a zombie apocalypse breaking out or my wife demanding my attention.

The solution is my TR2 multi-sport waterproof bone conduction headphones from H2O Audio (**h2oaudio.com**). These little beauties boast excellent sound quality, and they feature a built-in MP3 player with 8GB of memory that can store thousands of songs. Alternatively, I can stream anything I like from my iPhone using Bluetooth.

Now I can listen to music while still being able to hear and respond to my wife's dulcet demands. I wonder what the early audio pioneers would have thought of that! **PE**

Seaside Sound Simulator

Line Out
SILICON CHIP
JUST SURFIN'
12VDC
@ 150mA
CON1
LM324
BC549
2xBC337

By John Clarke

Relax and enjoy the sound of the seaside from the comfort of your home. Forget the scorching heat in summer, the cold winds of winter or your swimsuit being full of sand! Ideal for beginners and experienced constructors alike, it's a fun-filled project.

Image Source: https://unsplash.com/photos/birds-eyeview-of-seashore-3P3NHLZGCp8

Our new Seaside Sound Simulator uses standard through-hole components that mount on a blue PCB shaped like a surfboard.

It produces a sound that imitates the ebb and flow of the waves rolling up on the beach, including the occasional big wave. It can be used to augment the sound of surf if you live near the sea, or allow you to experience the beach even if you live in Birmingham. The sound is ideal for masking background noises so that you remain relaxed or for a peaceful sleep.

The project uses all standard parts and has a fun surfboard shape that includes graphics depicting waves. It includes an onboard loudspeaker, or you can use the RCA socket to feed the sound to a stereo system or powered speaker for an even more realistic effect. Using large speakers with extra bass will reproduce the deep thumps as the waves crash onto the beach.

It's powered by a 12V DC plugpack, so you don't have to worry about batteries going flat. It requires no adjustments to work. All you do is switch it on, set the volume and you're instantly drifting off, imagining a day at the beach.

Producing the wave sound

The sound of waves crashing on the shore is very similar to white noise, a randomly produced sound that covers the audio spectrum from 20Hz-20kHz (for humans). White noise has the same intensity level at every frequency. It is similar to the sound coming from an AM radio when it is not tuned to a radio station, or the noise produced by heavy rainfall.

Pure white noise needs some changes to sound like the surf. The volume needs to change over time and there needs to be some tailoring of the frequency response to sound realistic. There also must be some randomness to the waves since there is considerable variation in the sea noise as waves come into and crash onto the beach, then withdraw.

The volume levels of the waves have a triangular shape over time with some extra details. As a wave comes in, the sound steadily increases, hits its peak and then dies away. To simulate wave sounds, we use a white noise source that has its volume varied by triangular 'envelopes'. By having two such envelopes, we can obtain a degree of randomness to the sound level.

With one generator, you only get the same wave crashing at a constant rate, but with two, you get two sets of waves rolling in at more unpredictable intervals. With further shaping of the triangular envelope, we can obtain extra wave sound realism.

This design is based on a circuit from October 1990 by Darren Yates that was published in Silicon Chip magazine. We have kept it based around two low-cost LM324 quad op amps; while we could have reduced the component count using a microcontroller, that would have been less interesting and harder to modify.

This version features some improvements to the circuit and it is considerably more compact and appealing on the surfboard-shaped PCB rather than in a plastic box.

Block diagram

Fig.1 shows the block diagram of the Seaside Sound Simulator circuitry. The preamplifier, IC2c, provides the main sound output. It is fed white noise to its non-inverting (+) input, while the volume (or amplifier gain) is altered over time using two triangle wave generators and three modulators, designated MOD1, MOD2 and MOD3. The modulators change the shape of the triangular envelopes.

The output of triangle wave generator MOD1 is also fed to a peak amplifier, IC2d. This amplifies just the peak of the triangular waveform, where it increases the triangle wave output level. After feeding this voltage into another modulator (MOD3), it is used to produce a large wave crash simulation for when the wave hits the beach.

All three modulators vary the impedance from IC2c's inverting input to ground, changing the gain and therefore the sound level of the white noise.

The output of preamplifier IC2c is fed to a low-pass filter stage comprising IC2b and some passive components. This changes the frequency response of the white noise so that the higher frequencies are reduced, more like water sounds. From there, the signal is available at the CON2 line output for connection to an external amplifier and loudspeaker.

This signal is also fed to the volume control (VR1) for the power amplifier, IC2a, that drives the onboard loudspeaker.

Circuit details

Refer now to Fig.2, which shows all the circuit details. It is similar to the October 1990 version, with some variations. Some changes are simply because DC mains plugpacks these days are switch-mode types that provide a stable voltage under load, so we don't need a separate regulator.

In the 1990s, plugpacks generally comprised a mains transformer, bridge rectifier and filter capacitors. They provided a higher voltage with no load that dropped as current was drawn from the supply. The ripple also increased under load. For a voltage sensitive circuit, regulation was required.

Other changes were to isolate the supply between the sensitive circuitry used to produce the wave sound from the amplifier that drives the loudspeaker. This allows a higher volume level, as the 1990 version was a little too quiet. Without the isolation and with higher volume levels, the circuit would oscillate, producing a squealing noise as well as 'motor boating'.

While 'motor boating' might seem like a reasonable thing to include in a wave sound simulator, it is actually an electronic term to describe a low-frequency circuit oscillation malfunction. This is where a circuit produces its output in bursts, a bit like the put-put sound of a single-cylinder motor in a boat.

Another change was to prevent click and pop noises when parts of the circuitry suddenly change voltage level, from near 0V to near 12V or vice versa. We will describe those changes as we come to them in the following circuit description.

The main part of the circuit is the noise source. This is based on NPN transistor Q1. Its base-emitter junction is connected as a reverse-biased diode. This junction breaks down when the supply is in the reverse direction, allowing current to flow when the voltage across it reaches about 5V. The breakdown is a random process that produces considerable white noise.

To avoid damage, the current through the transistor junction is limited to around 200µA using the 33kΩ resistor to the +12V supply.

This noise is capacitively coupled to the non-inverting input (pin 10) of op amp stage IC2c. Two 100kΩ resistors connected in series across the 12V supply provide a 6V bias for IC2c so that its output can swing symmetrically within the 12V supply range.

Triangle wave generators

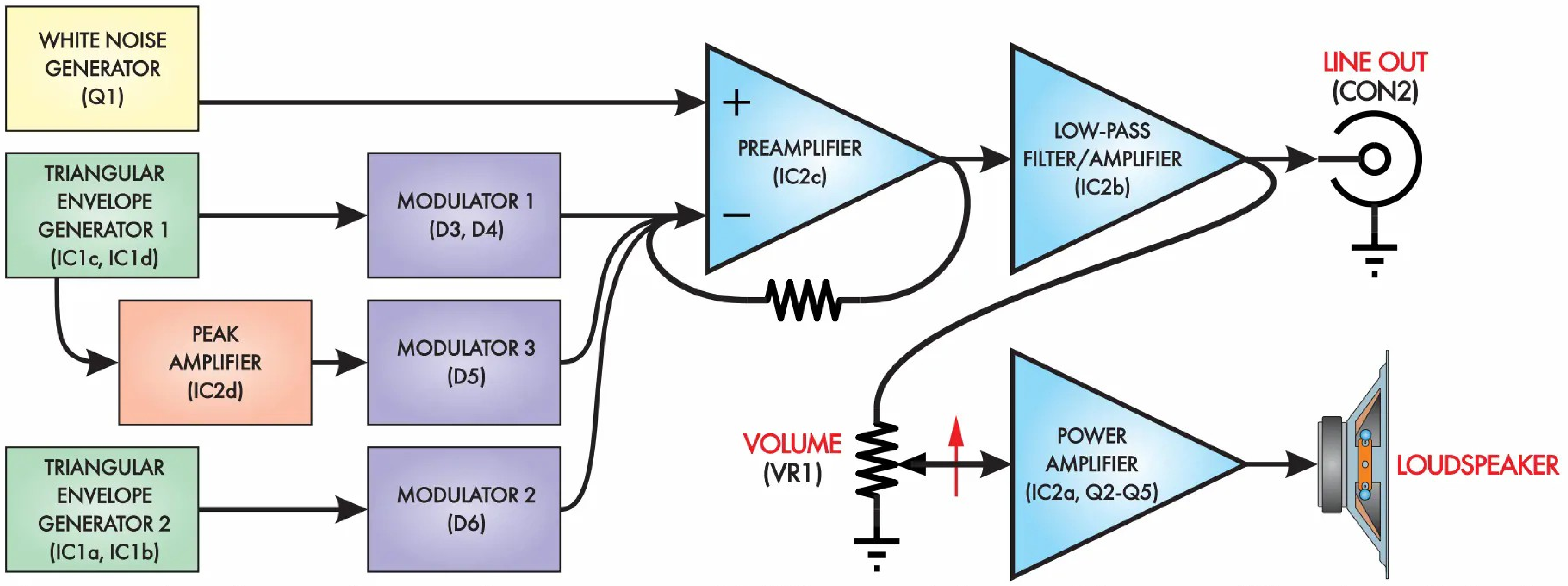
IC1d & IC1c together form the first triangle wave generator, while IC1a & IC1b form the second. The first generator is responsible for a wave that sounds very close (louder), while the second produces a wave that crashes in the distance (lower in volume). Because the two are nearly identical, we'll just describe how one of them works, then mention the slight differences between the two.

IC1d acts a Schmitt-trigger gate, while IC1c is connected as an integrator. IC1d's output will be either high (around 10.5V) or low (near 0V). It charges or discharges the 33µF capacitor at different rates depending on whether it is high or low. When the output is low, the capacitor charges via the 680kΩ resistor and series diode (D1) plus the parallel 330kΩ resistor. When IC1d's output is high, the capacitor charges only via the 330kΩ resistor.

The 33µF capacitor charge increases in a linear fashion toward the positive supply when the pin 8 output of IC1c is low, while it discharges linearly toward the 0V supply when that output goes high.

If you are interested in a more detailed (and complicated) description of how this works, see the panel titled "Triangle wave generation".

The only difference in the second triangle generator based on IC1a & IC1b is that the second generator has some lower-value resistors (100kΩ &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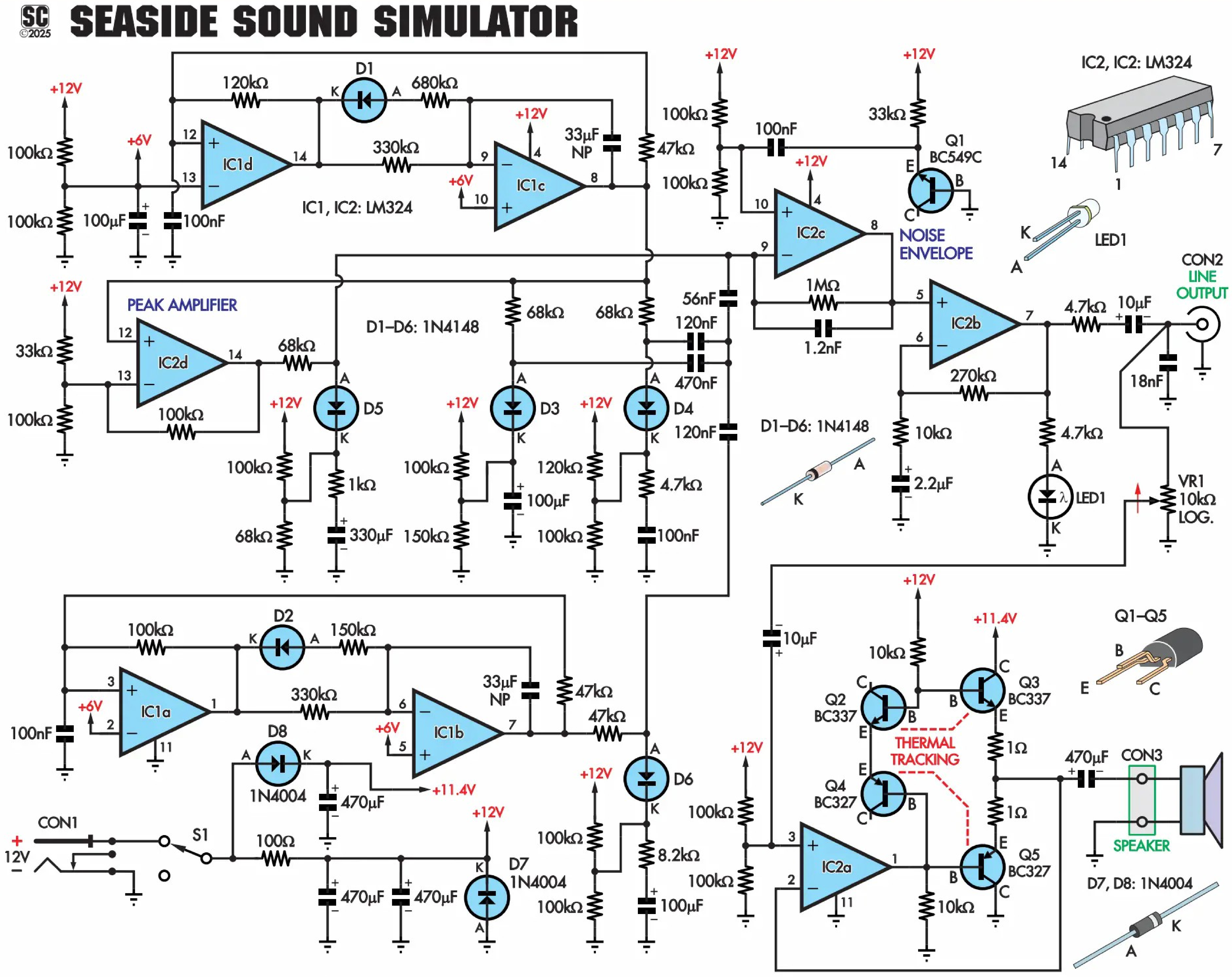
TRIANGULAR ENVELOPE GENERATOR 1 (IC1c, IC1d)
PEAK AMPLIFIER (IC2d)
TRIANGULAR ENVELOPE GENERATOR 2 (IC1a, IC1b)
MODULATOR 1 (D3, D4)
MODULATOR 3 (D5)
MODULATOR 2 (D6)
+
−
PREAMPLIFIER (IC2c)
LOW-PASS FILTER/AMPLIFIER (IC2b)
LINE OUT (CON2)
VOLUME (VR1)
POWER AMPLIFIER (IC2a, Q2-Q5)
LOUDSPEAKER

Fig.1: two triangular waveform envelope generators with different periods control the preamplifier gain applied to the white noise source. Three modulators and one peak amplifier tweak the sound to make it more like waves crashing on the shore. The resulting audio is filtered and fed to the line output (CON2) plus a volume control (VR1) and power amplifier to drive the onboard loudspeaker.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14
IC1d
330kΩ
9
10
8
4
IC1c
33μF NP
47kΩ
100μF
100nF
IC1, IC2: LM324
100kΩ
33kΩ
100nF
Q1 BC549C
E
B
C
10
9
IC2c
NOISE ENVELOPE
IC2, IC2: LM324
14
7
1
LED1
K
A
CON2 LINE OUTPUT
PEAK AMPLIFIER
IC2d
68kΩ
100kΩ
D1–D6: 1N4148
68kΩ
68kΩ
56nF
1MΩ
120nF
1.2nF
470nF
5
6
7
IC2b
4.7kΩ
10μF
18nF
270kΩ
D5
D3
D4
1kΩ
330μF
150kΩ
120kΩ
100μF
4.7kΩ
100nF
120nF
10kΩ
2.2μF
4.7kΩ
LED1
VR1 10kΩ LOG.
D2
100kΩ
150kΩ
3
1
2
11
IC1a
330kΩ
6
5
7
IC1b
33μF NP
47kΩ
47kΩ
100nF
D8
1N4004
470μF
+11.4V
D6
8.2kΩ
100μF
10μF
10kΩ
Q2 BC337
Q3 BC337
THERMAL TRACKING
Q4 BC327
Q5 BC327
1Ω
1Ω
470μF
CON3
SPEAKER
IC2a
10kΩ
Q1–Q5
B
E
C
CON1
12V
S1
100Ω
470μF
470μF
D7 1N4004
D7, D8: 1N4004
K
A

Fig.2: it helps to refer to the block diagram, Fig.1, when trying to understand how this circuit works. Transistors Q2 & Q4, diode-connected in series, produce a bias voltage for current buffer transistors Q3 & Q5 that tracks over temperature, to avoid thermal runaway.

150kΩ instead of 120kΩ & 680kΩ). It helps to make the two waves more random because the two generators run at different speeds. It also provides the second wave with a faster 'travel rate' towards the shore.

One of the problems with the triangle generators is that the Schmitt trigger outputs (pins 1 & 14) produce a clicking sound whenever the voltage from their output swings between 0V and 10.5V. The 1990 circuit used 100nF capacitors from the outputs to ground to suppress this, but on building the circuit in 2024, we found it wasn't that effective.

Without any capacitors, the rise time of those outputs was 25μs; with the capacitors, it was reduced to 18μs, worsening it! We found that placing the 100nF capacitors at the non-inverting inputs of the op amps, at pin 12 for IC1d and pin 3 for IC1a, significantly increased the output rise time to 75μs. The clicks and pops went away.

There are still two clicks that occur when the Seaside Sound Simulator is initially switched on, but no more are evident after that.

Diode modulators

The outputs of the two triangle wave generators drive the diode modulator circuits as shown in the block diagram (Fig.1). These rely on the fact that the conductivity of a diode varies with the voltage across it, ie, a diode with 0.6V across it will conduct more current than one with only 0.2V across it.

There are three modulators in the circuit, based on diodes D3 to D6. Diodes D3 & D4 connect to the same IC1c output, so are counted as one modulator. The first triangle generator drives D3 & D4, the second drives D6, while the third (D5) modulator is driven by the peak amplifier, IC2d.

At the cathodes of these diodes is a voltage divider. In the case of D6, for example, there is a pair of 100kΩ resistors. These set the offset voltage for this modulator to 6V. Different resistance values are used in the voltage dividers of the other modulators. These set the offset levels to different values to ensure the correct switch-on sequence.

For diode D6, this means that the output of its triangle wave generator must rise above 6V before the diode has enough forward bias to conduct. This output is coupled to the anode of D6 via a 47kΩ resistor and also to the inverting input of preamplifier IC2c via a 120nF capacitor.

While the voltage from IC1b remains below 6V, D6 is reverse-biased and the 120nF capacitor sees a high-impedance to ground. However, when the voltage rises above 6V, the diode begins to conduct, which decreases its AC impedance. The 120nF capacitor thus sees a progressively lower impedance to ground as the voltage across the diode increases.

Since op amp IC2c is connected as a non-inverting amplifier, these impedance variations directly control its gain. If the impedance goes down, the gain goes up and vice versa. Thus, the diode modulators control the gain of the preamplifier stage to vary the sound level.

When the voltage across D6 reaches 0.6V, the diode appears as a short-circuit to the capacitor and the impedance to ground is then set by the 8.2kΩ resistor connected to D6's cathode. The 100μF capacitor and 8.2kΩ resistor form a high-pass filter that rolls off the response below 0.2Hz.

D3 and D4 work similarly but have offset voltages of 7.2V and 5.45V, respectively. Note also that D4 controls another high-pass filter, consisting of a 4.7kΩ resistor and 100nF capacitor, with a -3dB point of 340Hz. Because of their different offset voltages, D4 comes into operation before D3 (which controls lower frequencies), so we get a realistic "whoooosshhh" sound as the wave breaks.

Peak amplifier

The gain of IC2c is also controlled by diode modulator D5, which is driven by peak amplifier IC2d. Its input comes from the output of IC1c. The bias for IC2d's inverting input (pin 13) is set to about 7V by the 33kΩ resistor and the two 100kΩ resistors. Thus, the output of IC2d remains low until pin 8 of IC1c reaches this threshold level.

At that point, IC2d amplifies the signal to produce a faster, steeper waveform. This produces the big 'dumper' sound of a wave that crashes onto the beach.

Low-pass filtering

As IC2c amplifies the white noise generated by Q1, a 1.2nF capacitor in the feedback loop of IC2c rolls off the response above 130Hz. The 2.2μF capacitor in the feedback network of IC2b also rolls off the low frequency response of this stage below 7Hz.

IC2b is a non-inverting amplifier with

Triangle wave generation

For the first triangle wave generator, IC1d forms a Schmitt trigger gate, while IC1c is connected as an integrator. IC1d's output will be either high (around 10.5V) or low (near 0V) with different charge and discharge rates for the 33μF capacitor.

The capacitor charges when IC1d's output is low via the 680kΩ resistor and series diode, plus the parallel 330kΩ resistor, but only discharges via the 330kΩ resistor when IC1d's output is high.

The IC1d Schmitt trigger receives the voltage from IC1c's pin 8 via a 47kΩ resistor to the non-inverting input (pin 12). Hysteresis is provided by the 120kΩ resistor from pin 12 to IC1d's output. When IC1d's output is low, pin 12 input is pulled lower via the 120kΩ resistor and the voltage divider formed with the 47kΩ resistor that monitors the IC1c output. The inverting input at pin 13 is at 6V.

IC1d's output will go low when pin 12 rises above the pin 13 voltage. Knowing that pin 14 of IC1d is low (0V) and that pin 8 is rising, we can find the voltage where pin 8 causes pin 12 to be at 6V.

When there is 6V across the 120kΩ resistor, 50μA flows through it. The voltage at pin 8 must be sufficient to produce a 50μA flow through the 47kΩ resistor that has its pin 12 end at 6V. The voltage across the 47kΩ resistor will be 2.35V (47kΩ x 50μA). So pin 8 would be 8.35V (6V + 2.35V).

This means that the 33μF capacitor charges to 8.35V at pin 8. The pin 9 side remains at 6V as IC1c adjusts its output to maintain this 6V.

With pin 12 of IC1d just above 6V, its output goes high to around 10.5V. Now the capacitor (and pin 8 of IC1c) begins to discharge toward 0V via the 330kΩ resistor. Diode D1 is reversed-biased in this case.

We can calculate what the pin 8 voltage will be when pin 12 just falls to 6V again. Since IC1d's output is at 10.5V and pin 12 will be at 6V, the voltage across the 120kΩ resistor will be 4.5V (10.5 – 6V). So the current through the 120kΩ resistor will be 37.5μA. This same current flow is through the 47kΩ resistor, so it will have 1.76V across it, below 6V, giving 4.23V.

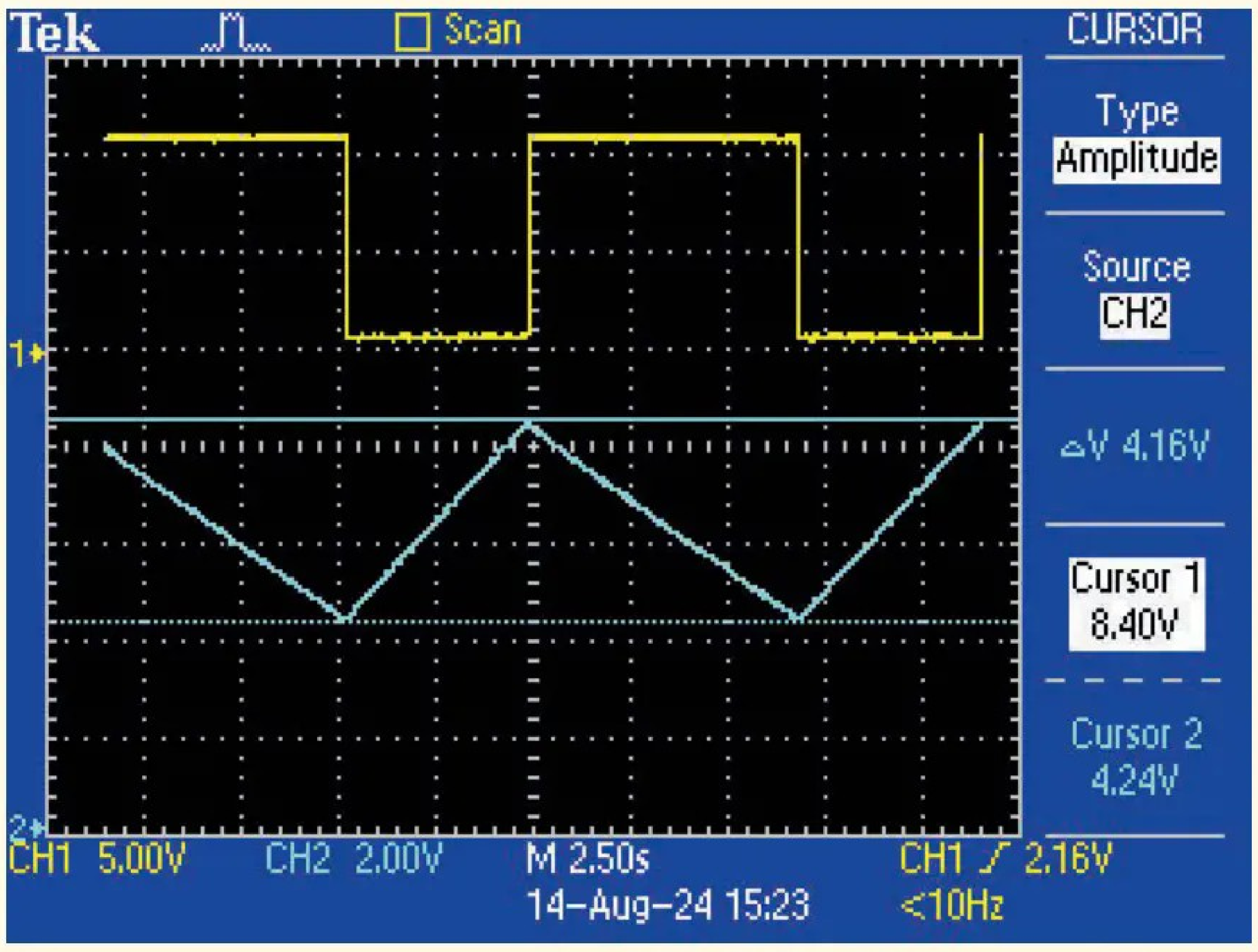
Once this voltage is reached, the output of IC1d drops again to recharge the capacitor in the positive direction. We ignore any current to the non-inverting input of the op amp, as that will be just 100nA at most.

As the two switching levels are 4.5V and 8.3V, that means there is a 3.8V hysteresis provided by the 120kΩ resistor. Without this, there would be no controlled oscillation.

The resulting waveform at pin 8 of IC1c will be a sawtooth, a triangular shape rising faster than it falls. Partly this is because the LM324's output can pull down to nearly 0V but can only go up to about 10.5V when powered from 12V. The other reason is that there is an extra current path via D1 when IC1d's output is low.

Scope 1 shows oscilloscope traces of IC1d's output (pin 8) in the top yellow waveform and the triangle waveform output from pin 8 of IC1c in cyan. The triangle wave swings between 4.2V and 8.4V, close to the values calculated above.

The faster charge and slower fall time for the triangle wave has the overall effect matches the sound of ocean waves, which come up to shore faster than they run back to the sea.



Scope 1: the triangular sawtooth waveform generated at pin 8 of IC1c is shown in the lower (cyan) trace. On the right, the 'scope indicates that the voltage difference between the peak and trough is 4.16V. The voltage at pin 8 of IC1d that dictates whether the capacitor is being charged (low) or discharged (high) is shown above in yellow.

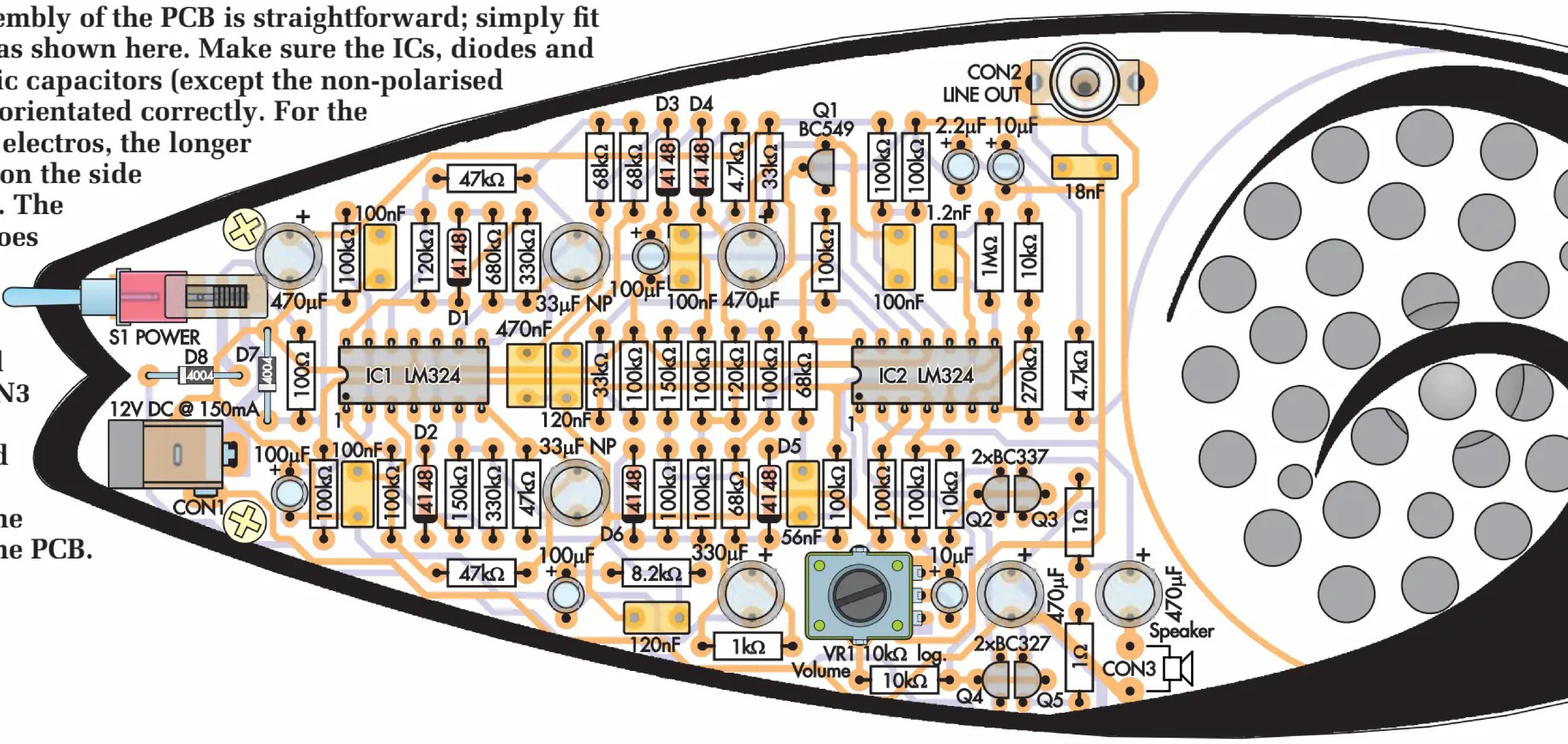

CON2
LINE OUT
D3 D4
Q1
BC549
2.2µF 10µF
18nF
47kΩ
68kΩ
68kΩ
4148
4148
4.7kΩ
33kΩ
100kΩ
100kΩ
1.2nF
100nF
100kΩ
120kΩ
4148
680kΩ
330kΩ
100µF
100nF 470µF
100kΩ
100nF
1MΩ
10kΩ
470µF
33µF NP
D1
470nF
S1 POWER
D8
4004
D7
4004
100Ω
IC1 LM324
33kΩ
100kΩ
150kΩ
100kΩ
120kΩ
100kΩ
68kΩ
IC2 LM324
270kΩ
4.7kΩ
12V DC @ 150mA
120nF
33µF NP
D2
D5
100µF
100nF
100kΩ
100kΩ
4148
150kΩ
330kΩ
47kΩ
4148
100kΩ
100kΩ
68kΩ
4148
100kΩ
100kΩ
100kΩ
10kΩ
2xBC337
Q2
Q3
1Ω
CON1
D6
56nF
47kΩ
100µF
8.2kΩ
330µF
10µF
470µF
470µF
Speaker
120nF
1kΩ
VR1 10kΩ log.
Volume
10kΩ
2xBC327
Q4
Q5
1Ω
CON3

Fig.3: assembly of the PCB is straightforward; simply fit the parts as shown here. Make sure the ICs, diodes and electrolytic capacitors (except the non-polarised ones) are orientated correctly. For the polarised electros, the longer lead goes on the side marked +. The speaker goes on the rear of the PCB; it is wired to the CON3 terminals and sound passes through the holes in the PCB.

a gain of 28. The original 1990 circuit used a gain of 11 for this amplifier, but with supply routing changes (described later), a higher gain is possible. It is a significant increase in the maximum volume at just over 8dB.

Indicator LED1 is driven from the IC2b amplifier output via a 4.7kΩ resistor. The LED will light with varying brightness and, to some extent, mimic the sound level.

Following IC2b is another low-pass filter stage comprising a 4.7kΩ resistor, a 10µF coupling capacitor and an 18nF filter capacitor. The 18nF capacitor rolls off the response above 1.88kHz to reduce higher frequencies further, adding realism to the sound.

After that, the signal goes to the CON2 RCA socket and also to the 10kΩ volume control pot (VR1), which feeds the signal to the power amplifier, based on op amp IC2a and transistors Q2 to Q5. Q3 and Q5 buffer the output of the op amp to provide current gain; they are within IC2a's feedback loop to reduce crossover distortion.

Transistors Q2 and Q4 produce a bias voltage for the output transistors (Q3 & Q5). Only the base (B) and emitter (E) terminals of these transistors are connected, using them as diodes to produce a nominal 0.6V bias. These diode junctions match the voltage across the output transistor (Q3 and Q5) base-emitter junctions.

The bias voltage ensures the output transistors are always conducting current and this reduces crossover distortion as signal swing passes the mid (6V) level where the output drive switches between Q3 and Q5.

This type of amplifier is called Class-AB. Class-B means that one output transistor conducts for positive excursions and the other conducts for negative excursions. It also has some amount of Class-A operation at low signal levels, where both Q3 and Q5 are conducting, due to a small standing current through both transistors at these low levels.

The 1Ω emitter resistors provide a degree of bias current stability. A higher bias current will cause extra voltage across the 1Ω resistors that effectively raises the bias required for Q3 and Q5 to conduct, reducing the current through them. The bias voltages from Q2 and Q4 remain more-or-less constant unless the temperature changes.

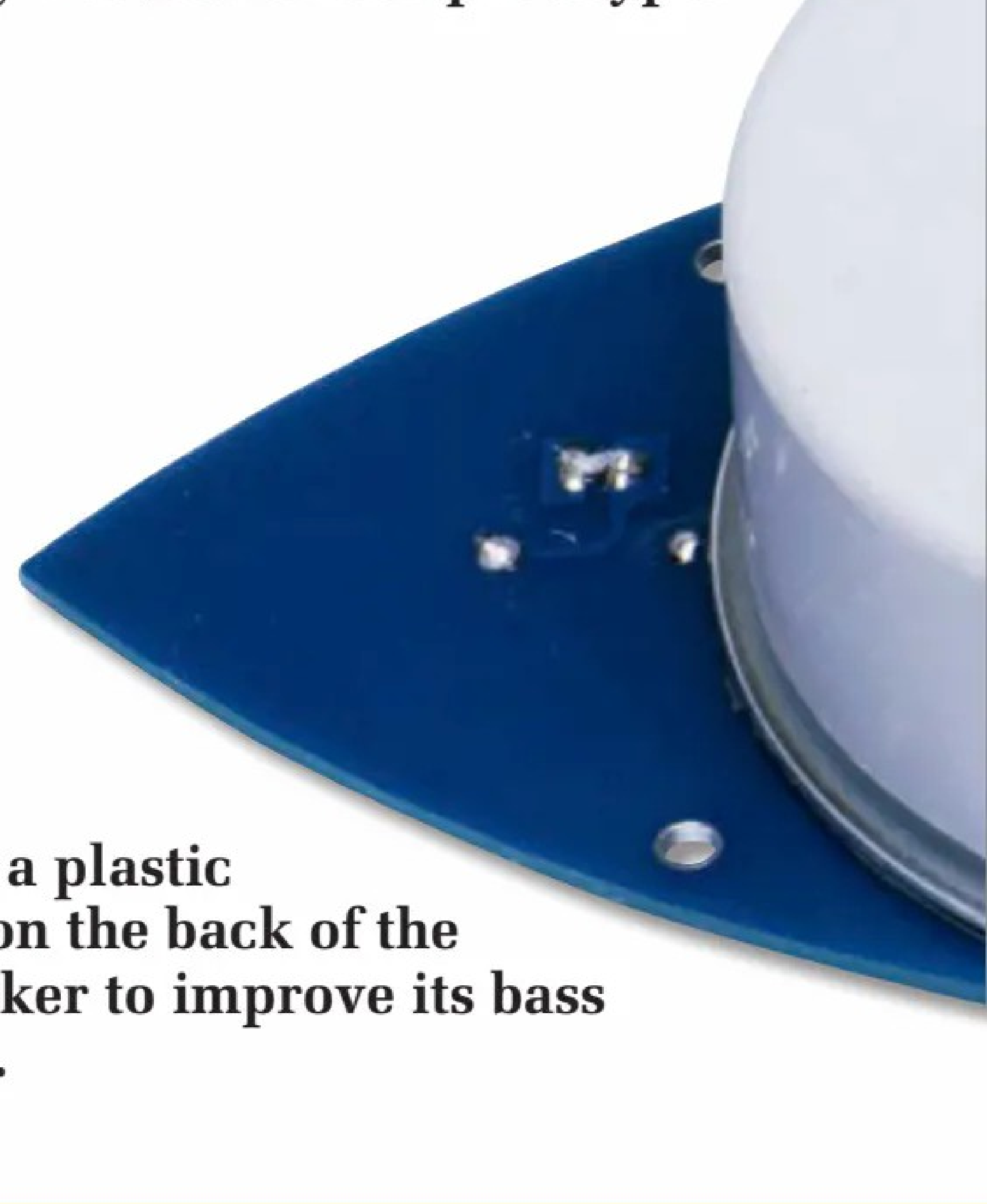
The bias current is kept steady with temperature because Q2 is physically touching Q3 and Q4 is touching Q5, so the transistor pairs maintain a similar temperature. This prevents thermal runaway should the output transistors heat up when driving a load like a loudspeaker.

Without the thermal matching and with a fixed bias, as Q3 and Q5 heat up and their base-emitter voltages drop, the current through them would increase, causing more heating and thus thermal runaway. In our circuit, Q2 and Q4 will reduce the bias voltage as they heat up, preventing that.

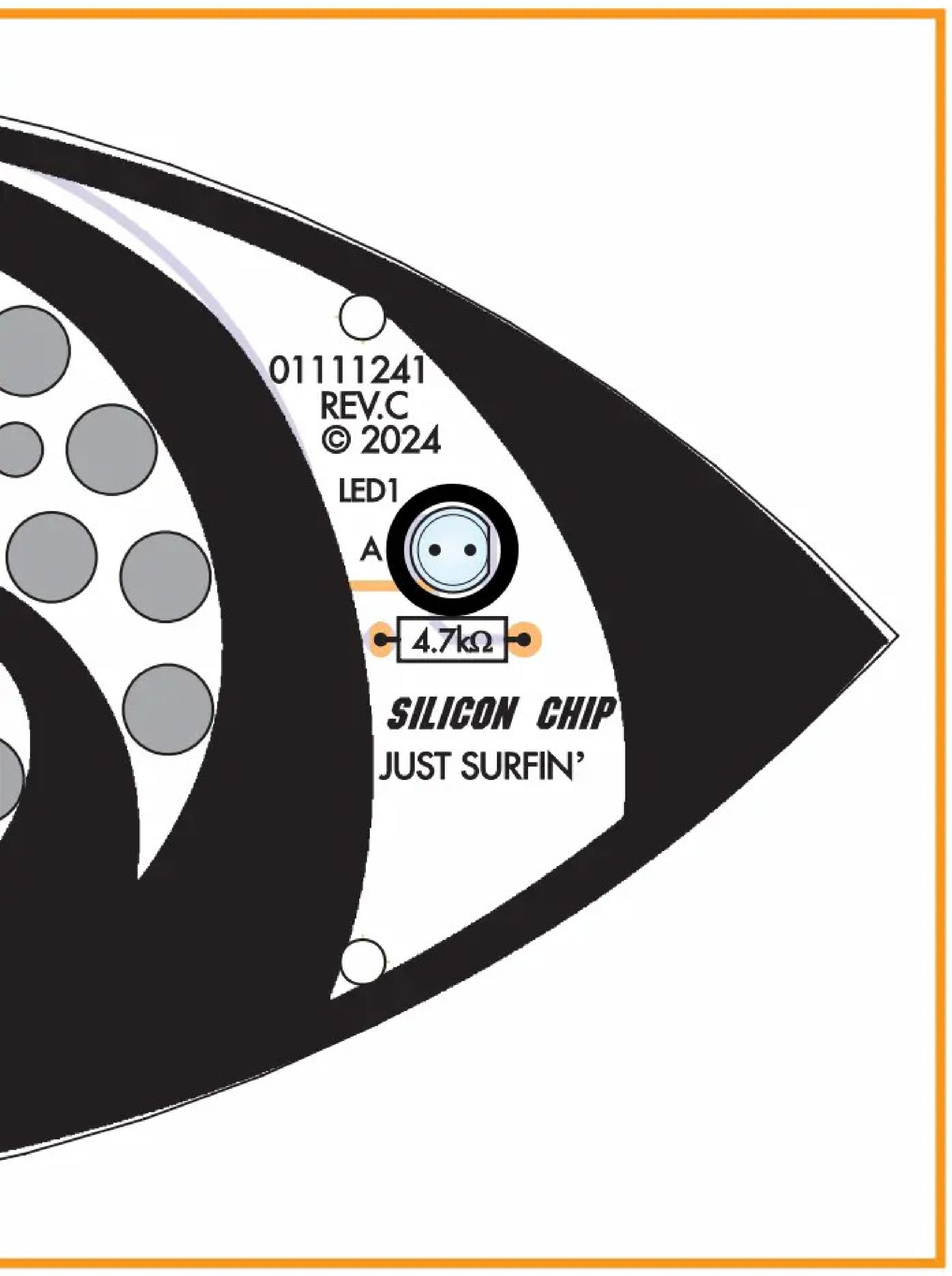
Q3 and Q5 drive the loudspeaker via the 1Ω resistors and the 470µF coupling capacitor. This capacitor removes the 6V DC offset of the amplifier so that the loudspeaker is driven purely by an AC voltage.

Power for the circuit is from a 12V DC plugpack connected to CON1. Switch S1 connects this supply via two paths. One path is via the 100Ω resistor to power the op amps. This supply is bypassed using two 470µF capacitors. The second path is via diode D8 to the loudspeaker amplifier circuitry, bypassed by another 470µF capacitor.

Note that apart from the two wires for the loudspeaker, the other components that you can see fitted to this side are not required, and were only needed for our prototype.



We have installed a plastic end cap on the back of the loudspeaker to improve its bass response.



The 100Ω isolation resistor prevents the circuit from oscillating and motor boating, as mentioned previously. This resistor, along with the two 470μF capacitors, maintains a stable voltage for the op amp circuitry that is separate from the loudspeaker driver supply.

Without this isolation, any supply voltage change due to current drawn to drive the loudspeaker would reduce the op amp supply voltage, causing the wave sound generator voltages to vary, leading to motor boating.

Reverse supply polarity protection is provided by diodes D7 and D8. D7 clamps the reverse voltage applied to the circuit to -0.6V; the current through it is limited to 114mA by the 100Ω resistor. D8 provides reverse polarity protection for the 470μF capacitor that bypasses the loudspeaker driver supply. It prevents any current from flowing if the supply polarity is wrong.

Construction

All components for the Seaside Sound Simulator mount on a double-sided blue PCB coded 01111241 that measures 236 × 80mm.

As shown on the overlay diagram, Fig.3, most parts are on the top of the PCB. Only the loudspeaker is on the other side. An end cap is attached to the rear of the loudspeaker to improve its bass response.

Begin by fitting the resistors. The colour codes for these are shown in the parts list but it is best also to check the values using a multimeter. Some of the colours can be difficult to discern against the blue background body colour of the resistor.

Install the diodes next. D1-D6 are the smaller 1N4148 types, while D7 and D8 are 1N4004s. Take care to fit each with the correct orientation. The two IC sockets can be mounted next. Again, these need to be orientated correctly, with the notched section at the end with pins 1 & 14 as shown.

The MKT polyester capacitors can now go in. These are not polarised, so they can go either way around. They will likely be marked with a code rather than the actual value; the likely codes are shown in the parts list.

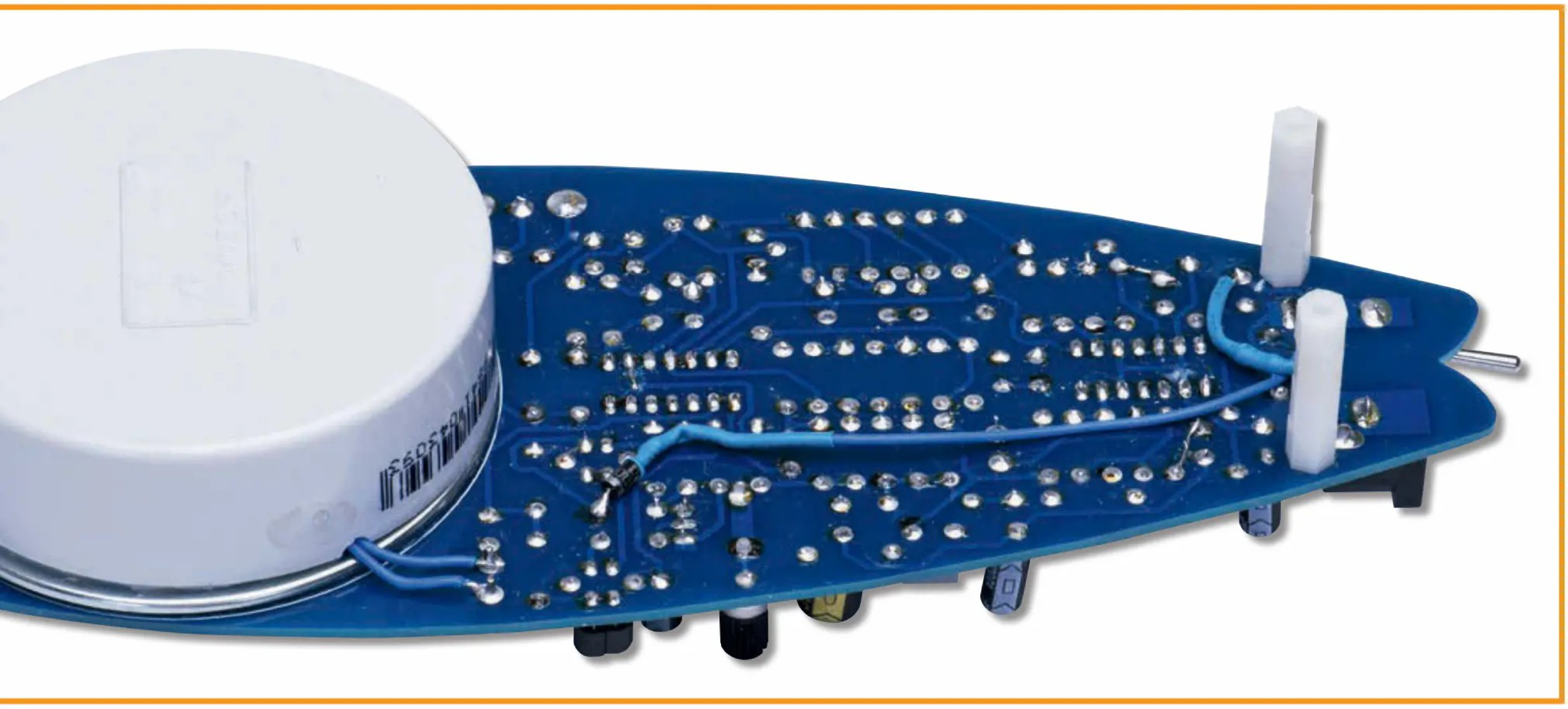
The transistors can go in next. There are three types: a BC549C for Q1, BC337s for Q2 & Q3 and BC327s for Q4 & Q5. Take care to install each in the correct position. Q2/Q3 and Q4/Q5 have their flat sides facing each other. Ideally, they should touch each other (perhaps with a smear of thermal paste between) so their temperatures track.

Install CON1 (the PCB-mounting DC barrel socket) and switch S1 now. S1 can be a toggle or slider switch, as specified in the parts list (there are likely other suitable parts). LED1 can also be fitted now; ensure it goes in with the anode (longer lead) in the hole marked "A". It can sit close to the PCB.

The electrolytic capacitors are next. Most of these are polarised, so they must be orientated with the correct polarity. The plus sign (+) on the PCB shows the positive side, which corresponds to the longer capacitor lead. The striped side of the can is the opposite (negative) side. The two 33μF capacitors are non-polarised (NP) types, so they can be mounted either way.

Now install CON2 (the PCB-mounting RCA socket) and potentiometer VR1. Insert and solder two PC stakes (PCB pins) at the CON3 speaker connection points.

The loudspeaker mounts on the back of the PCB and is connected to those stakes/pins with two short lengths of hookup wire. For the



moment, the loudspeaker can be left off the PCB and connected with wire leads for testing.

Testing

Insert the two LM324 ICs into their sockets with the pin 1 and notched end orientated correctly. Make sure that when you push them down, the pins go into the socket and don't get folded up under them.

When you power the unit up with a 12V DC supply and S1 on, you should see LED1 light and hear the wave sound coming from the loudspeaker. If not, check the setting of the volume potentiometer, VR1.

If there is still no sound, check the supplies to IC1 and IC2. There should be around 11.75V between their pins 11 and 4. Also check your construction for correct component locations and orientations.

Once you are satisfied with the operation, the loudspeaker can be secured to the rear of the PCB using neutral-cure silicone sealant (roof & gutter sealant), contact adhesive or any other suitable glue. A 76mm-diameter screen-printed circle is provided on the back of the PCB to show the ideal position.

We attached a PVC plumbing end cap (a 65mm DWV [Drain Waste and Vent] type) to the rear of the loudspeaker to provide a baffle for it, giving extra bass. A small notch will need to be made with a round file to allow the speaker wires to enter the bottom edge of the end cap. The end cap can then be secured to the rear of the loudspeaker with the same glue used for the speaker

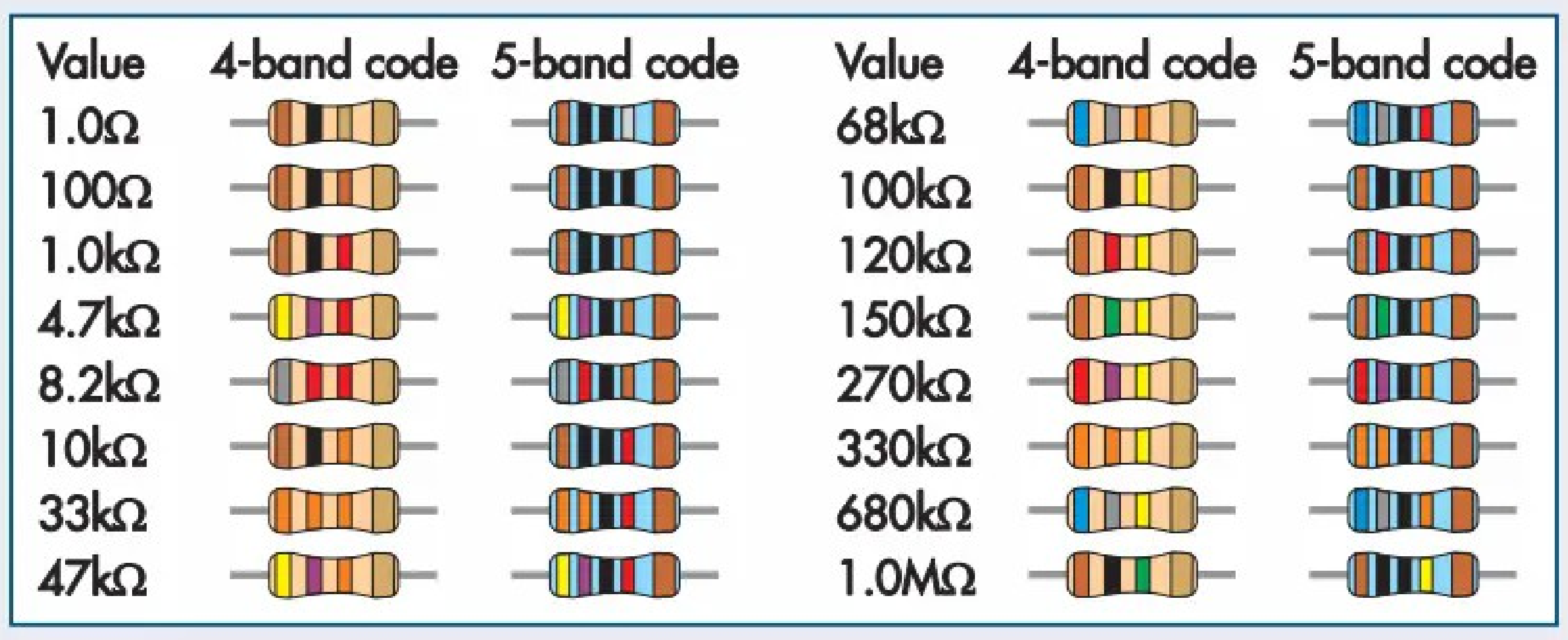
M3 screws and spacers can be attached at the PCB mounting hole near CON1 and switch S1 so that the Seaside Sound Simulator can sit horizontally or lean back vertically on the plumbing fitting at the rear of the loudspeaker and the lower standoff. **PE**

Parts List – Seaside Sound Simulator

1 double-sided plated-through blue PCB coded 01111241, 236 × 80mm
1 76mm (3in) 8Ω 1W loudspeaker (SPK1) **[Farnell 599256]**
1 12V DC 150mA+ plugpack
1 PCB-mount DC barrel socket to suit plugpack (CON1)
1 vertical PCB-mounting RCA socket (CON2) **[Farnell 2817597]**
1 10kΩ logarithmic vertical 9mm PCB-mounting potentiometer (VR1) **[Farnell 2519595]**
1 PCB-mounting 90° SPDT toggle or vertical slide switch (S1) **[Farnell 9574638]**
2 14-pin DIL IC sockets
2 1mm diameter PC stakes (for CON3)
1 63mm ID, 70mm OD (2.5in) uPVC DWV end cap **[Amazon UK B0DLWKC71Z]**
2 M3-tapped 25-30mm Nylon standoffs or spacers, length to suit end cap (use 10+15mm or 15+15mm if you can't get the required lengths)
2 M3 × 25mm panhead machine screws

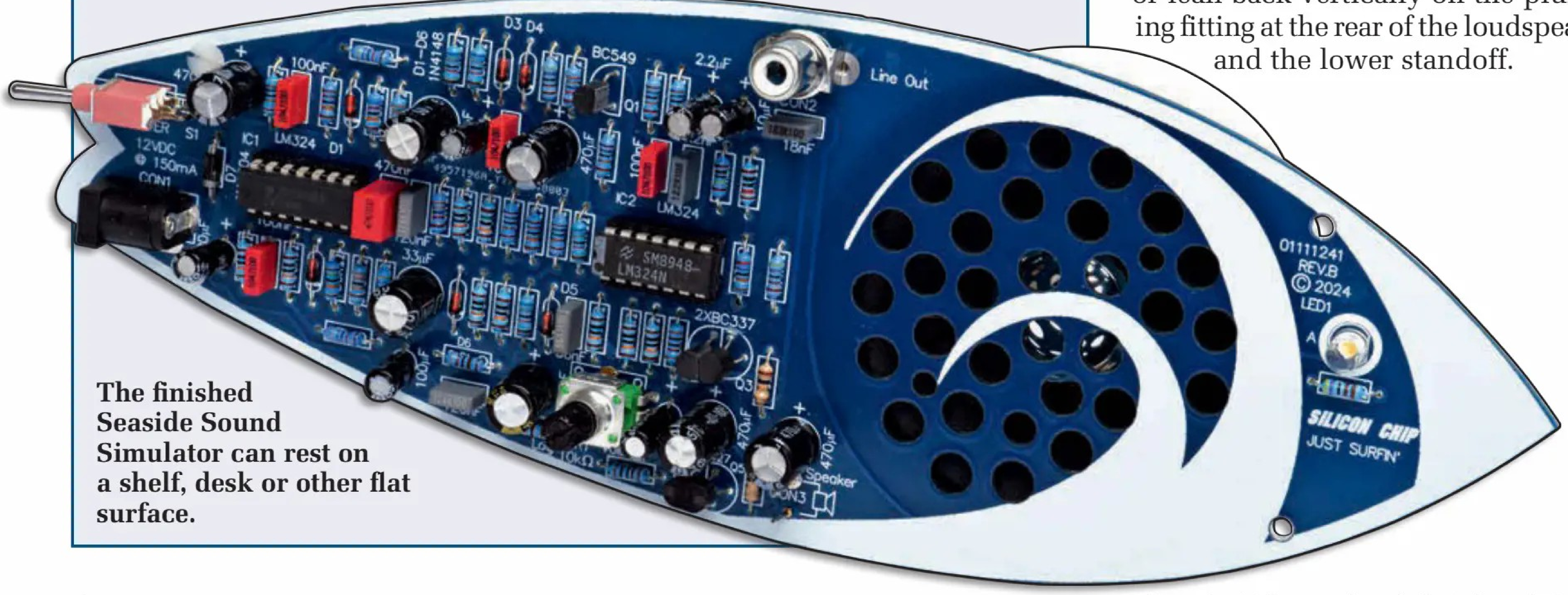
Semiconductors
2 LM324 quad single-supply op amps (IC1, IC2)
1 BC549 (ideally BC549C) 30V 100mA NPN transistor (Q1)
2 BC337 45V 800mA NPN transistors (Q2, Q3)
2 BC327 45V 800mA PNP transistors (Q4, Q5)
6 1N4148 75V 200mA diodes (D1-D6)
2 1N4004 400V 1A diodes (D7, D8)
1 5mm white LED (LED1)

Capacitors *(all 63/100V MKT polyester unless noted)*
4 470μF 16V PC electrolytic
1 330μF 16V PC electrolytic
3 100μF 16V PC electrolytic
2 33μF 50V NP (non-polarised) PC electrolytic
2 10μF 16V PC electrolytic
1 2.2μF 63V PC electrolytic
1 470nF (code 474)
2 120nF (code 124)
4 100nF (code 104)
1 56nF (code 563)
1 18nF (code 183)
1 1.2nF (code 122)



Value	4-band code	5-band code	Value	4-band code	5-band code
1.0Ω			68kΩ		
100Ω			100kΩ		
1.0kΩ			120kΩ		
4.7kΩ			150kΩ		
8.2kΩ			270kΩ		
10kΩ			330kΩ		
33kΩ			680kΩ		
47kΩ			1.0MΩ		

Resistors *(all axial ¼W 1%)*
1 1MΩ
1 680kΩ
2 330kΩ
1 270kΩ
2 150kΩ
2 120kΩ
14 100kΩ
4 68kΩ
3 47kΩ
2 33kΩ
3 10kΩ
1 8.2kΩ
3 4.7kΩ
1 1kΩ
1 100Ω
2 1Ω 5%



The finished Seaside Sound Simulator can rest on a shelf, desk or other flat surface.

AOShop

The home for specialist audio, analogue and historic components – provided by Jake Rothman, *PE's Audio Out* columnist. The AOShop is your best bet for classic analogue 'NOS' (new old stock) components, including all parts for *Audio Out* projects and designs.

Ge semiconductors

Small-signal PNP transistors

NKT214F, OC57, OC59, 2N1377, 2N525	**£1.00**

Low-noise PNP transistors

GET106	**£2.50**

Small power PNP transistors

AC153, AC153K, AC188	**£1.50**

Small power NPN transistors

AC176, AC141K	**£2.00**

High-voltage PNP transistors

OC77, CV7001	**£1.50**

TO3 PNP power transistors

OC22, CV7054 (OC23), OC25, OC35, OC36, AD143, AD149, AD161, AD162	**£2.00**
AD140	**£3.50**
AD149, AD161/2 matched pair	**£5.00**

RF PNP transistors

OC41, OC42, 2SA12, 2SA53,AF124, AF178, GET872A	**£1.50**

NPN transistors

OC139, OC140, ASY73	**£2.00**

Diodes

CV7049 (OA10), CG92 (OA91)	**£0.50**

Si semiconductors

Diodes

ZC5800 RF Schottky	**£0.20**

Low-noise silicon transistors

BFW16A, 2SC3071, 2SC3068, 2SA1016K, 2SC2362K, 2SA970BL	**£1.50**
2SC2204, 2SD655, BC550C	**£0.50**
ZTX651	**£0.30**
ZTX751	**£0.50**

RF transistors (suitable for Theremin)

BF199	**£0.50**

Audio power MOSFETs

Exicon 10N20, 10P20	**£6.50**
Hitachi 2SJ99, 2SK343	**£3.50**
Hitachi 2SJ56, 2SK176	**£8.50**

JFETs

BFW11, BFW10, TIS73L, J177, J113, U1994, U1898, 2SJ176, J201	**£1.00**
J175, J176, J112, J111, 2N3820, 2N5467, BF244, 2N5460, J230	**£0.60**

Small power output/driver transistors

2SB649A, D669A, 2SA1208, 2SC2910, MJE253G, MJE243G, 2SA1725, 2SC4511	**£1.20**
BD139,BD140, BD135, BD156, BD435, BD436, MPSA63, BCV46	**£0.50**

MOSFETs

ZVP2106A	**£0.30**
Dual-gate 3SK45, BFS28	**£1.50**

Metal-cased transistors

BCY71	**£0.30**

Amplifiers

LM318 high-speed op amp	**£0.35**
µA709 metal-cased op amp	**£2.00**
5534H metal-cased op amp	**£3.00**
LM384, TDA2030A, TDA2050V	**£1.50**
TAA435 (Mullard power amp driver for AD161/AD162)	**£1.20**

Synthesiser ICs

That/dbx 2180 VCA/VCF	**£6.00**
CA3080 VCA/VCF (vocoder)	**£2.80**
CA3280 VCA/VCF	**£4.00**
LM13600/ LM13700 VCA/VCF	**£2.00**
CA3086 transistor array	**£1.00**
PT2399 echo/delay	**£1.50**

Dual transistors

2N2639, 2N2223, 2N2910 (NPN)	**£4.00**
E401 (JFET Moog)	**£4.00**
2SK2145-Y dual JFET	**£0.80**
2N5564 JFET	**£8.00**
HN3C51F, HN3A51F	**£1.00**
DMMT3904/6, HN1A01F, HN1C01F	**£0.50**

Loudspeakers

PE Mini-Monitor

Volt PE165 6.5-inch woofer (each)	**£85**
Morel MDT29 tweeter (each)	**£25**
Kit *pair* of PE165/MDT29 plus Volt crossover parts and PCBs	**£299**
Monacor DT-28N tweeter (each)	**£35**
Vifa 19mm BC20SC15-04 tweeter (each)	**£15**

Volt crossover inductors

1.2mH, 1.5mH, 2mH, 2.7mH, 0.5mH (tapped at 0.3mH) (each)	**£5.85**

Fully assembled and tested high-quality speaker prototypes – ask for details

LS3/5As and other similar speaker systems (pair)	**£200-£350**

Fully tested reclaimed speakers

Vifa BC14 5-inch woofer (each)	**£10**
Vifa TC26 1-inch tweeter (each)	**£10**

Low-price speakers

Philips 4-inch 4070	**£2.00**
EMI 10x6-inch, 30Ω Alnico	**£7**
64mm 64Ω neodymium	**£1.20**

Capacitors

Note '10/63' denotes '10µF 63V'.

Polyester

3.3/100, 4.7/250, 4.7/63	**£1.00**
5.6/63, 8.2/63, 10/63	**£2.00**
Mullard 'Mustard' C296 0.22/400	**£2.00**

Polycarbonate

Axial 2.2/63 1%, 4.7/160, 6.8/63	**£1.00**

Radial

6.8/160V, 10/63	**£2.00**
Reclaimed 22/63	**£2.00**

Polystyrene

Philips 1% 4.7nF/160, 6.2nF/500, 12nF/63, 22nF/63, 110nF/63, 24nF, 2nF	**£1.00**
RIFA 1% 100nF/100, Suflex 90.9nF 0.5%	**£2.00**
Suflex 2.5% 10nF/63 (rad. or ax.)	**£0.50**

Radiation resistant

Siemens cellulose acetate MKL 2.2/25	**£0.80**

Electrolytic – Mullard blue 017 series

10/25, 22/25, 100/10	**£0.50**
150/40, 470/40, 1000/40	**£1.00**

Tantalum – axial metal cased

22/50, 47/35, 68/25, 100/20, 120/10, 150/16, 220/10, 330/6	**£2.00**
22/35, 33/35, 47/20, 68/15, 100/10, 150/6	**£1.25**
4.7/50, 6.8/35, 10/25, 10/35, 22/15	**£1.00**

Axial moulded-case tantalum

Kemet axial 6.8/10	**£0.30**
Kemet radial 33/10	**£0.40**
STC radial 100/20	**£1.50**

Tantalum bead

22/50	**£1.00**
470/3	**£2.00**
680/6.3	**£3.00**

Wet tantalum

220/25 axial	**£3.20**
Castanet button 140/30, 470/3	**£3.20**
Hughes 540/10	**£3.20**

Bipolar

Hermetic bipolar tantalum 16/35	**£3.20**
Elcap axial 10/50	**£0.50**
Generic radial 100/16, 470/35, 100/ 63, 22/35, 4.7/35, 220/16	**£0.50**

Philips solid-aluminium (axial) 121/123

47/16	**£1.00**
330/6.3	**£2.00**
100/35	**£4.00**

Philips Pearl 122 series (radial)

10/16	**£0.30**

Silvered mica (radial)

1nF/500 1%	**£1.00**

Trimmer capacitors

Vishay plastic-film 4-40pF	**£0.80**
Vishay plastic-film 5.5-45pF	**£1.00**
Vishay plastic-film 5-80pF	**£1.20**
Vishay plastic-film 10-250pF	**£2.00**
Mica 1-12pF, 2-40pF	**£1.00**

Audio transformers and inductors

Eagle transformers

LT44, LT722 driver, LT700, LT723 500Ω output	**£2.50**
LT30 500mW output	**£3.50**
5:1 interstage (fuzz box)	**£1.50**

Repanco

T/T3 splitter transformer	**£4.00**
CH2 5mH RFC	**£2.00**

Balanced output transformer

Vigortronix 600Ω VTX-101-007	**£10**
Vigortronix 600Ω VTX-101-3001	**£10**
Vigortronix 600Ω VTX-101-3002	**£15**
Gardners 150Ω	**£10**
Reclaimed BBC LL74/MPC nickel core 600Ω	**£12**
Low-level 1:1 600Ω (modem)	**£2**
Reclaimed mic input transformer	**£15**

Inductors

82µH, 4.7mH, 100µH, 270µH, 10µH, 14µH (low Z)	**£0.50**
7-inch ferrite rod with MW and LW windings	**£3.00**

Special resistors

Bourns wire-wound trimmer 5kΩ 3059 JM panel-mount	**£2.00**
Thermistor RA53, R13	**£4.00**
A13	**£2.00**
Thermistor CZ1, CZ6	**£1.50**
Holco H2 2.2MΩ 1W, 1%	**£1.00**
Welwyn 1GΩ 2W	**£1.00**
5k Bourns 3321H cermet trimmer	**£0.50**

Potentiometers

Bourns 81

25kΩ lin cermet	**£2.00**
25kΩ lin, 5kΩ lin cond. plastic	**£3.00**

Bourns 91

10k dual-gang lin or log	**£10.00**

Plessey moulded-track

5kΩ log with switch	**£3.00**
50kΩ A/log	**£3.00**
Mil 250kΩ lin dual	**£5.00**

Alpha

16mm 4.7kΩ A/log	**£0.80**
220kΩ A/log	**£0.80**
10kΩ lin centre-detent dual-gang	**£1.50**

Allen Bradley J series/Honeywell

10kΩ lin 1 million cycles	**£5.00**
Blore Edwards AB 45 dual 5kΩ A/log with switch	**£3.50**
Alps RK9 dual-gang 5kΩ RD law	**£4.00**
Alps 50k log tapped motor. stereo	**£5.00**
BI P260 500kΩ log conductive plastic 1 million cycles	**£2.00**
Colvern wire-wound 100kΩ or 50kΩ dual-gang 3W	**£5.00**
Mil Spec hermetic 10Ω	**£8.00**

Miscellaneous

PCBs

Dual JFET Buffer (*PE*, Oct '25)	**£2.00**
Mosfet Buffer (*PE*, Oct 2025)	**£2.00**
Dual Mosfet Buffer (*PE*, Oct '25)	**£2.00**
Cascode Amp (*PE*, Sep 2025)	**£2.00**
Cascode Power Supply (*PE*, Sep 2025)	**£2.00**
Pocket Theremin (*EPE*, 1996)	**£2.00**
Elysian Theremin (*EPE*, 1996)	**£6.00**
Synth VCF, VCO (*EPE*, 2017)	**£3.00**
48V PSU (*EPE*, 2019)	**£3.00**
SMT dual transistor adapter	**£0.40**

Relays

STC hermetic 4190 24V DPDT	**£4.00**

Switches Blore Edwards/Plessey

MH1 4-pole 3-way	**£3.00**
Elma 01 2-pole 6-way	**£6.00**
3PDT guitar pedal stomp switch	**£2.50**

Valves

EC88, PCL84	**£3.00**

Contact

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Shropshire SY7 9NN
(visit by appointment)
+44 (0)1597 829102
jrothman1962@gmail.com

Minimum order £5.00 inc post
Quantity discounts negotiable

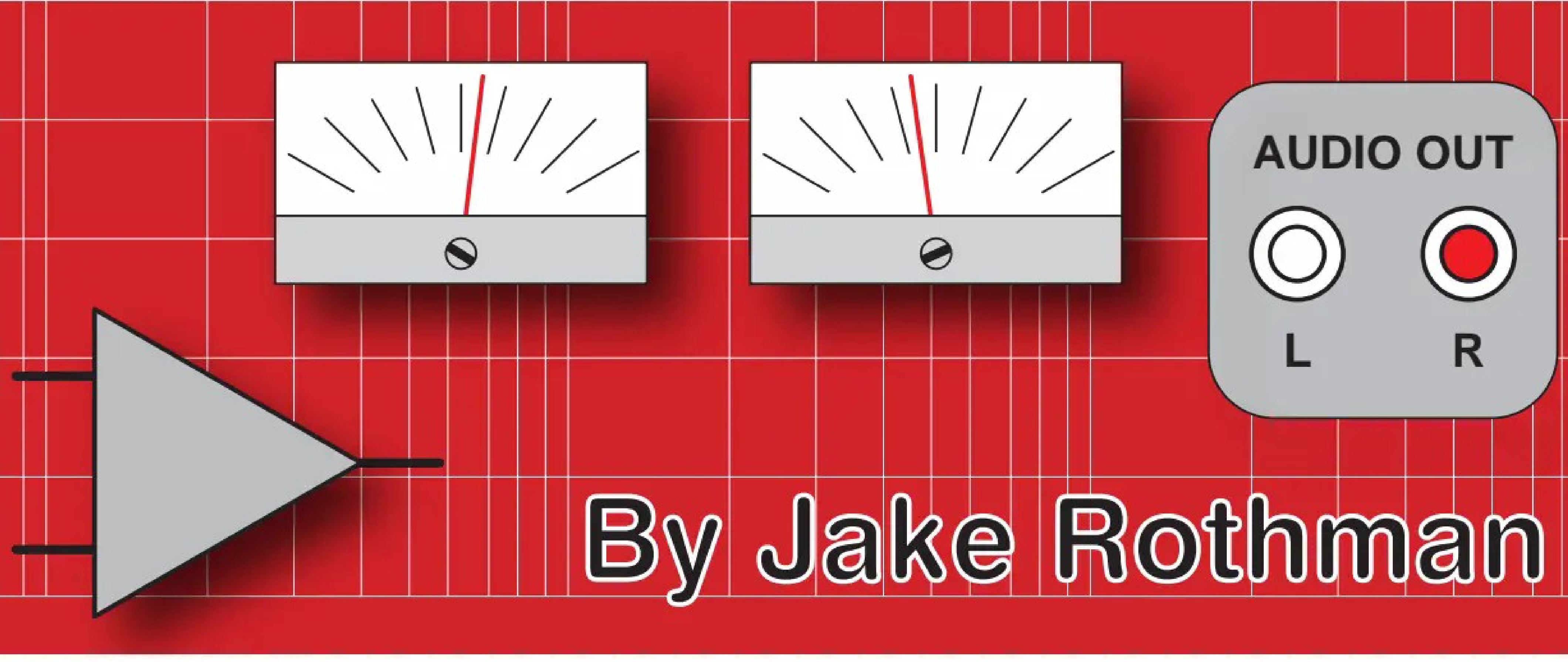
Payment

PayPal, bank transfer, cheques (payable to "J Rothman", UK pounds only)
No VAT payable

Postage

Small Jiffy bag £2.99
Small package £4.99
Big boxes and overseas at cost – ask for a quote

AUDIO OUT



By Jake Rothman

Three discrete class-A buffer modules

One of my long-term goals with Audio Out is to make a complete range of audio modules, so that any analog audio system can be made, whether it's a massive synthesiser or minimalist Hi-Fi amplifier. Once a great modular system comes together, it can be combined and made into a viable product.

I used to make modular systems from the late 1970s to the 1980s, in the 'golden age' of audio modules, before I could design at component level. If you can get access to the April 1977 issue of *Practical Electronics*, have a look at the advert on the inside front cover for Stirling Sound/Bi-Pre-Pak. It's where I got the SS125 modules, which I resurrected in the July 2025 issue.

There were at least nine audio module companies in that issue. My soldering iron just dripped with nostalgic lust.

By the way, regarding that July issue, I've found a substitute for the ZTX694B VAS transistor (TR3): the 2SC2364. It is available from Profusion and AOShop at just 13p each in 100s. Just to be awkward, its collector is the centre pin, like most Japanese transistors.

Boards for old buffers

Back in the February and May 2024 issues, I described the circuits for some FET-based class-A discrete buffer circuits. The good news is I've now had Grindle, my partner in ~~crime~~ design, create some PCB designs for them; three altogether. A circuit without a PCB today is almost of academic interest only, especially if it includes surface-mounting parts.

My son questioned the point of making discrete circuits with SMT parts at all, since it detracts from long life and repairability, which is the main point of discrete circuits. For the circuits here, it's simply a matter of cost; SMT matched dual transistors cost around 30-50p each, while their hermetic leaded metal-can equivalents cost £5-12 each.

Besides, SMT is widely used now, and rework gear is available at low cost. It is not overly difficult to operate once you have a little practice. And there's no fundamental reason that an SMD should have a shorter life than a through-hole equivalent (they're just packaging differences, after all). In fact, part of the reason that SMD was developed in the first place was the desire for reliability by organisations like NASA (see page 2).

I will probably design a dual-outline board for those who have access to such gems as LM394 transistors and 2N5564 JFETs; I could always do a dual outline on the board. Grindle and I do our bit to make our SMT boards repairable; he designs our PCBs with the values and part numbers clearly visible alongside each component.

If that unmarked ceramic capacitor dies, the person repairing it at least stands a chance.

Plug-in 8-pin DIL module

This is designed to replace dual op amp chips for those desiring a minimalist discrete solution. It has less noise and high-frequency distortion than the ubiquitous JFET-input TL072 often employed as buffers in the Sallen-Key filters used in active loudspeakers.

Fig.1 shows the distortion of a good original Texas TL072. Normally, if one is to improve on the TL072, a single-sourced and therefore expensive audio JFET input op amp has to be used, which is overkill for a buffer. My circuit, shown in Fig.2, is based on the White follower and uses a favourite device of mine, the Toshiba 2SK2145 dual JFET.

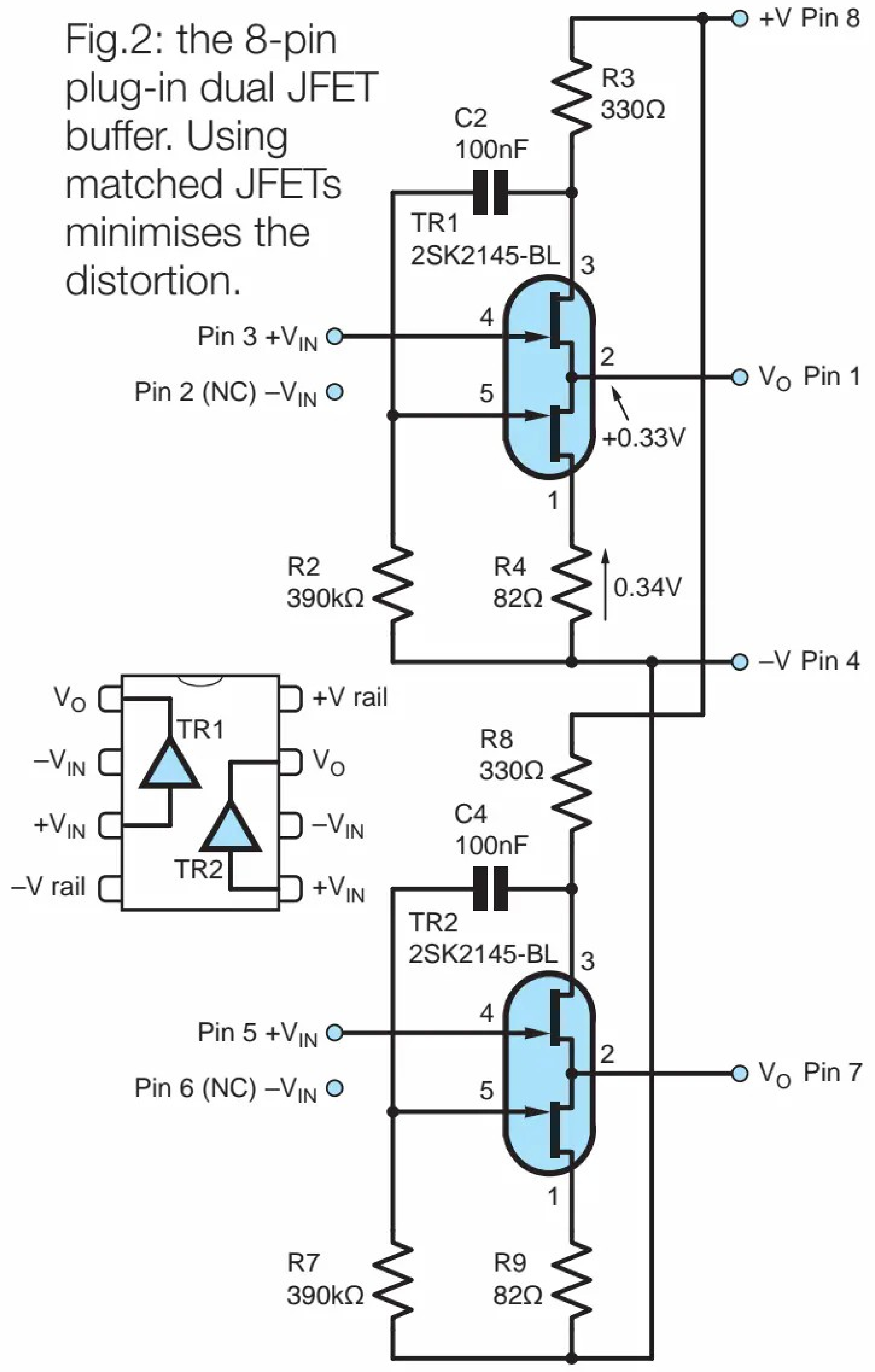


Fig.2: the 8-pin plug-in dual JFET buffer. Using matched JFETs minimises the distortion.

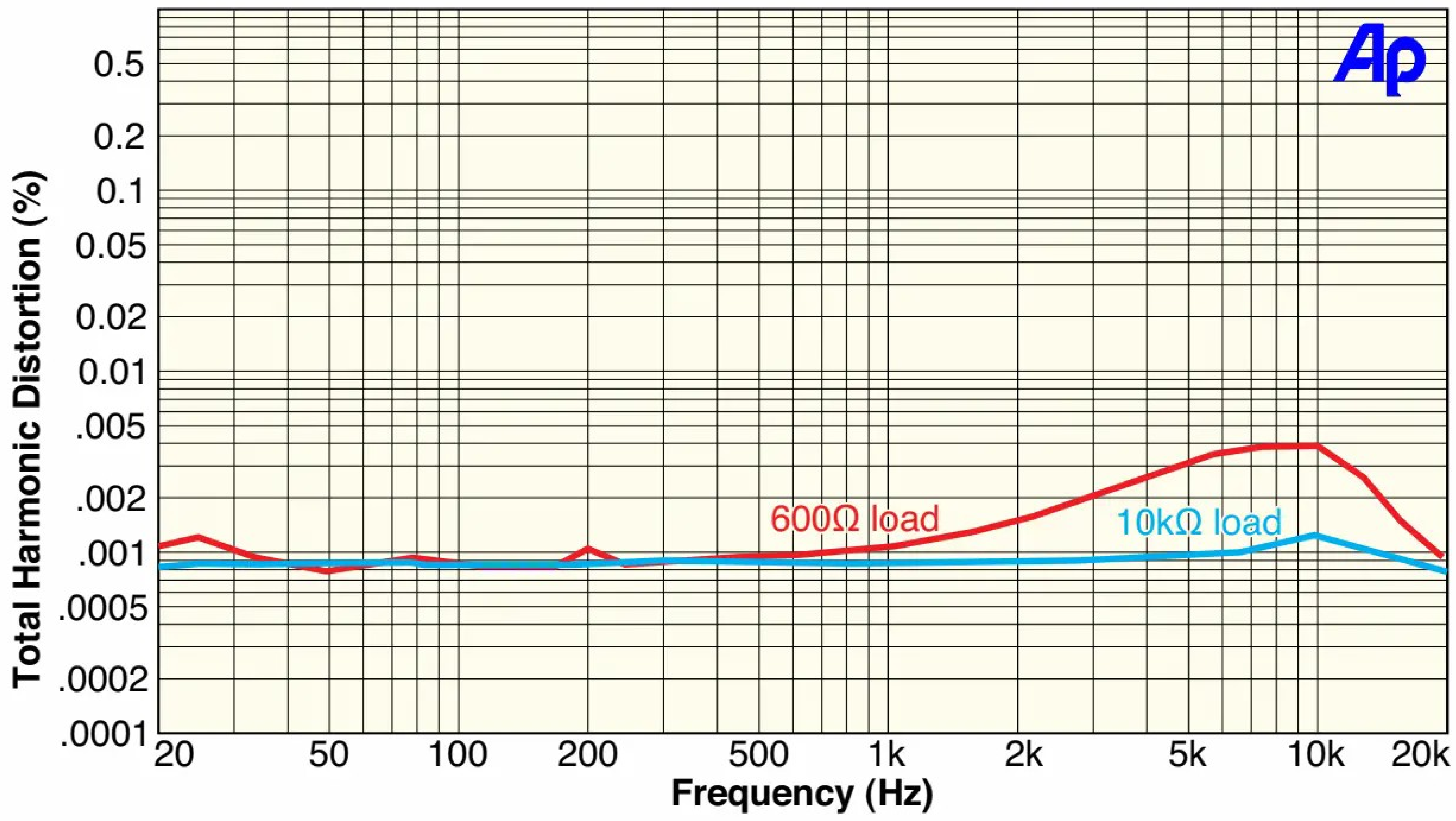


Fig.1: the distortion of a TI TL072 JFET-input op amp wired as a buffer at 1V RMS.

Parts List – Dual JFET buffer

1 double-sided 19 × 17mm PCB coded AO-OCT25-01
2 4-way 0.1-inch (2.54mm) pitch gold-plated round-pin headers (square pins won't fit into IC sockets) **[Tayda type X19525, SKU A-4662]**

Semiconductors
2 2SK2145-BL dual N-channel JFETs (TR1, TR2) **[Mouser 757-2SK2145BLTE85LF]**

Capacitors
2 100nF ±10% 63/100V polyester film, 5mm pitch (C2, C4)

Resistors (all SMD M3216/1206 ⅛W ±1% thin film)

2 390kΩ (R2, R7)	2 330Ω (R3, R8)	2 82Ω (R4, R9)

I made a little adaptor board for this device for experimenters, which was originally shown in the November 2023 issue of *Practical Electronics* on page 63 (Fig.56). The basic 2SK2145 specifications are:

- **Vgs maximum**: 50V
- **Transconductance/gain (Y_{FS})**: typically 15mA/V
- **Input capacitance**: 13pF
- **Noise factor (NF)**: 1dB @ 1kHz, source impedance 10kΩ
- **I_{DSS}**: 1.2-3mA (Y suffix), 2.6-6.5mA (GL suffix) or 6-14mA (BL suffix). The markings are X.L, XG and XL, respectively.
- **Power dissipation**: 300mW maximum (total for both FETs).

The BL grade is best for driving low impedances and has the lowest noise, while the Y is best for low-current battery applications, such as guitar pedals. Using the GL grade in the buffer circuit uses the same current as a TL072, in the order of 4mA (note the voltages across the resistors are halved). The specified BL grade draws 7.5mA, nearly that of the NE5532.

It has been in production since 1993, and like exceptional components by Toshiba valued by audio designers, it may be deleted soon, so I bought in a few hundred. Currently, they are 31p each in 100s from Mouser.

This design uses SMD resistors. Standard SMD resistors are mostly thick-film types that are non-linear and will introduce unwanted distortion, so make sure you stick to thin film types, which are available but a bit more expensive. They perform just as well as through-hole (axial) thin-film resistors. Having said that, standard thick-film types from Tayda work fine.

Specifications

The distortion curves for buffers with BL and GL JFETs are shown in Fig.3. Note that there is no rise at HF, a characteristic of op amps when used as buffers. This is because high levels of HF compensation are not required for stability at unity gain with the discrete buffers.

Since the JFET is rated at 50V, the buffer can be used at up to ±23V, but one has to be careful with loading to make sure the dissipation limit is not exceeded. The output impedance is higher than an op amp, at 100Ω.

Construction

Fig.4 shows the overlay, and the completed buffer is shown in Photo 1.

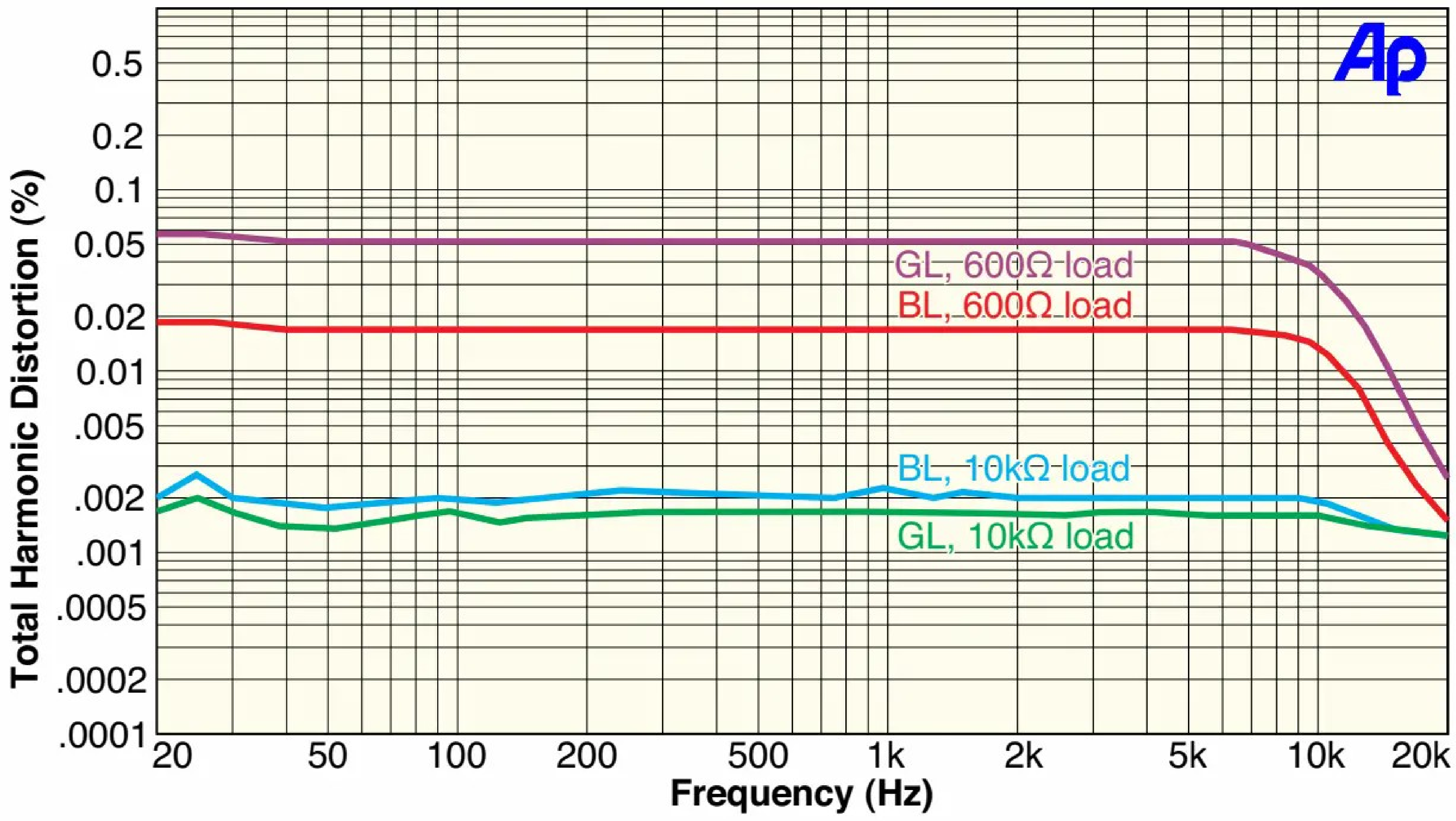


Fig.3: the distortion curve for the Fig.2 circuit using the BL and GL versions of the 2SK2145. The BL version is delivering 4V RMS into both 600Ω and 10kΩ, while the lower-current GL version is delivering 1V RMS into 600Ω & 4V RMS into 10kΩ.

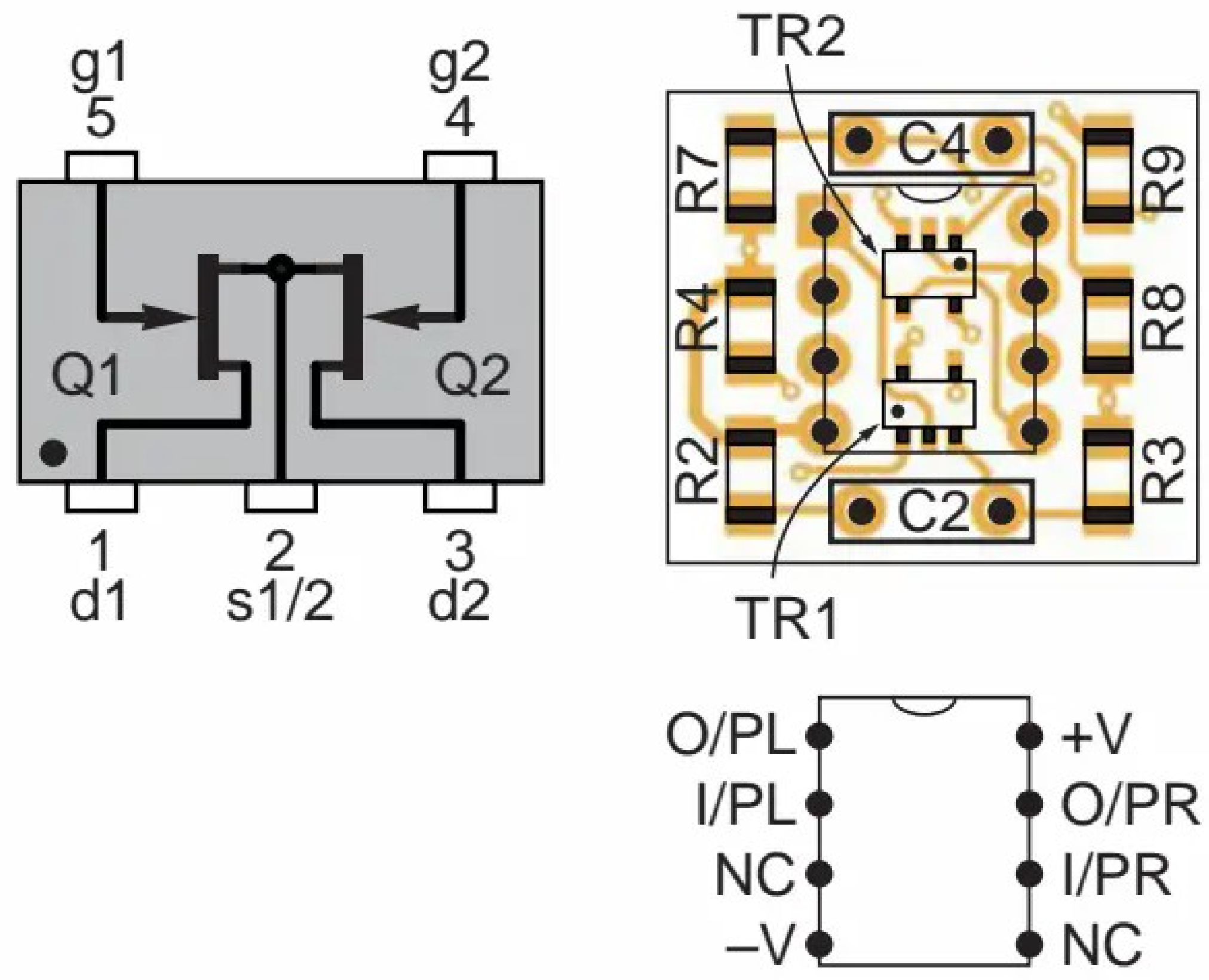


Fig.4: the dual JFET buffer component overlay. You can't get anything the wrong way around!

The SMD dual JFETs are considerably smaller than through-hole equivalents, so if your vision is far from perfect, you will want good light and a decent magnifier (ie, with an anti-reflection coating). A syringe of flux paste makes soldering the small leads much easier by allowing the solder to flow more readily.

Start by mounting the PCB so it doesn't move about. One way is to first put the IC socket pin strips into a breadboard and push the board on top to solder it. This way, it aligns the pins and provides a firm base at the same time (see Photos 2 & 3 overleaf) so you can solder them. Next, solder the JFETs in place. Luckily, they have asymmetrical packages, so you can't get them the wrong way around.

Add the resistors around the edge next, then add the radial plastic-film modulation capacitors, C2 and C4. Don't use ceramic X7R types as they will cause bad low-frequency distortion.

I made a version with AVX 100nF 35V SMT tantalum chip types, which gave low distortion. They had to be positioned with correct polarity, with the positive end facing the drain of the upper JFET. However, they cost ten times as much as the film types.

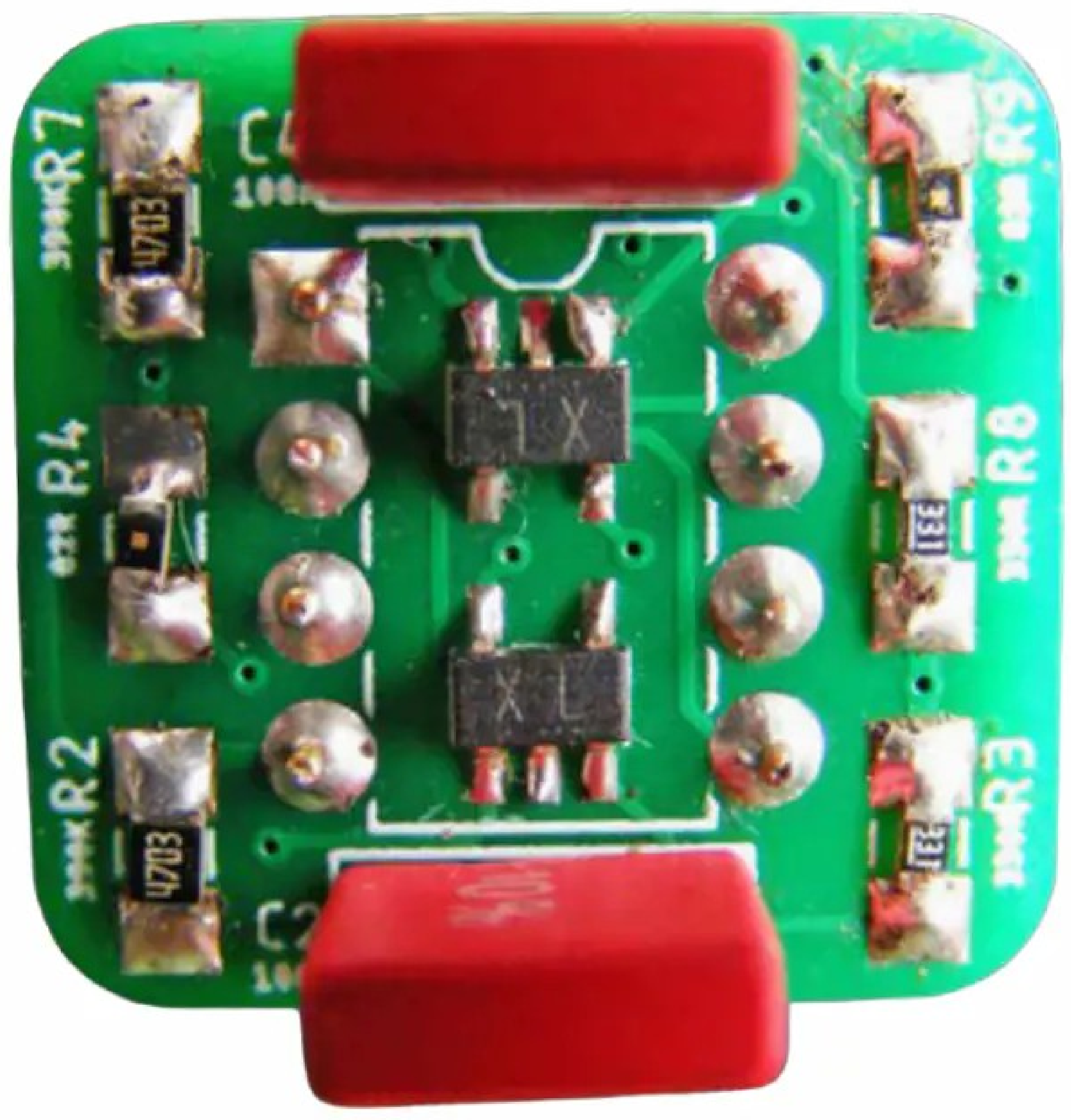
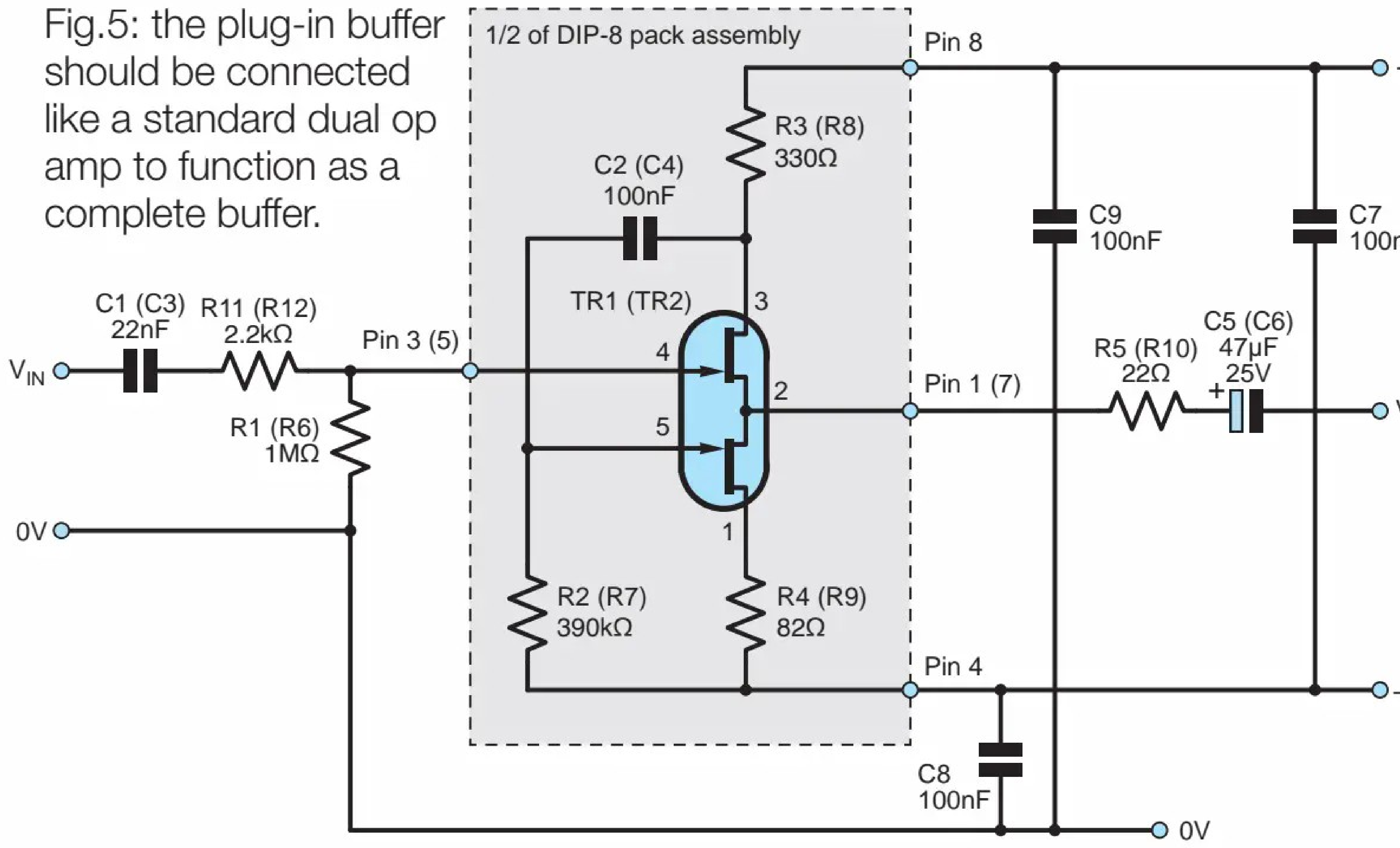


Photo 1: the discrete plug-in dual buffer. I used slightly smaller resistors than specified as I already had them.

Fig.5: the plug-in buffer should be connected like a standard dual op amp to function as a complete buffer.


1/2 of DIP-8 pack assembly
Pin 8
+V
R3 (R8)
330Ω
C2 (C4)
100nF
C9
100nF
C7
100nF
TR1 (TR2)
3
4
2
5
1
C1 (C3)
22nF
R11 (R12)
2.2kΩ
Pin 3 (5)
V_{IN}
R1 (R6)
1MΩ
0V
Pin 1 (7)
R5 (R10)
22Ω
C5 (C6)
47µF
25V
V_O
R2 (R7)
390kΩ
R4 (R9)
82Ω
Pin 4
–V
C8
100nF
0V

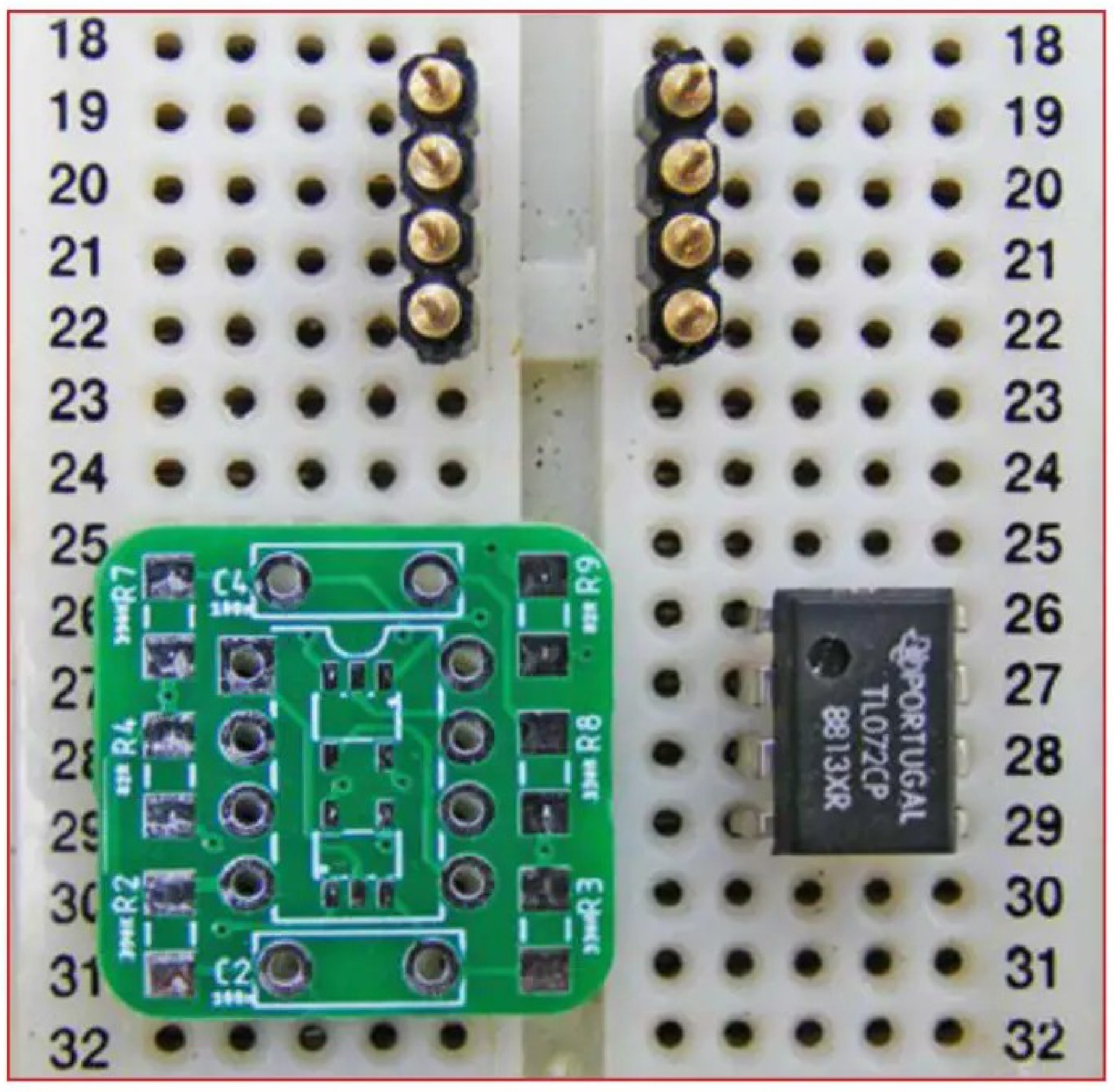

18
19
20
21
22
23
24
25
26
27
28
29
30
31
32
C4
R7
R9
R8
R2
C2
R3
PORTUGAL
TL072CP
8813XR

Photo 2: a breadboard can provide support to align the pins. The TL072 IC shows the relative size.

Soldering

Every engineer has their own take on SMT hand-soldering. I always like to coat the pads with a flux pen, then let them dry a bit to help stick the device in the correct position. Solder one pin quickly, before it has a chance to move. It helps if you use Multicore 26 SWG (0.46mm diameter) low melting point (LMP) solder. This is 36% lead, 62% tin and 2% silver, with 3% activated rosin flux.

If you use that awful low-residue 96% tin, 4% copper, lead-free stuff, expect trouble. Tin/lead solder is still used in military and avionic work. If you put too much solder on and bridge the pins, a bit of 3mm desolder braid usually helps soak it up (adding a little extra flux paste makes it work even better).

You can get rid of the flux residue with a good scrub with a stiff bristle artist's brush and isopropyl alcohol. That's good practice, even if you're using no-clean flux, as it can prevent you from seeing problems in the solder joints.

It's worth checking not only for possible bridges between adjacent pins on the JFETs but also to make sure that the solder fillet is covering both the pin/lead and the PCB pad. It's possible to get solder to stick to the lead but not the pad, which may not be obvious unless you look closely under magnification (with good light).

Testing

Recall that this unit is designed to replace a dual op amp used as a buffer, so it needs to be surrounded by normal buffer supporting circuitry to check that it works. A suitable test circuit is shown in Fig.5. There are pull-down resistors on the gates (R1, R6) along with gate resistors (R11, R12) to prevent oscillation. The same applies to the output resistors.

Capacitors C7, C8 & C9 are required for rail decoupling, with C5 & C6 for output DC offset blocking. Note that 'pins' 2 & 6 are not connected (NC). In practice, these may be linked directly to the output pins on most PCBs.

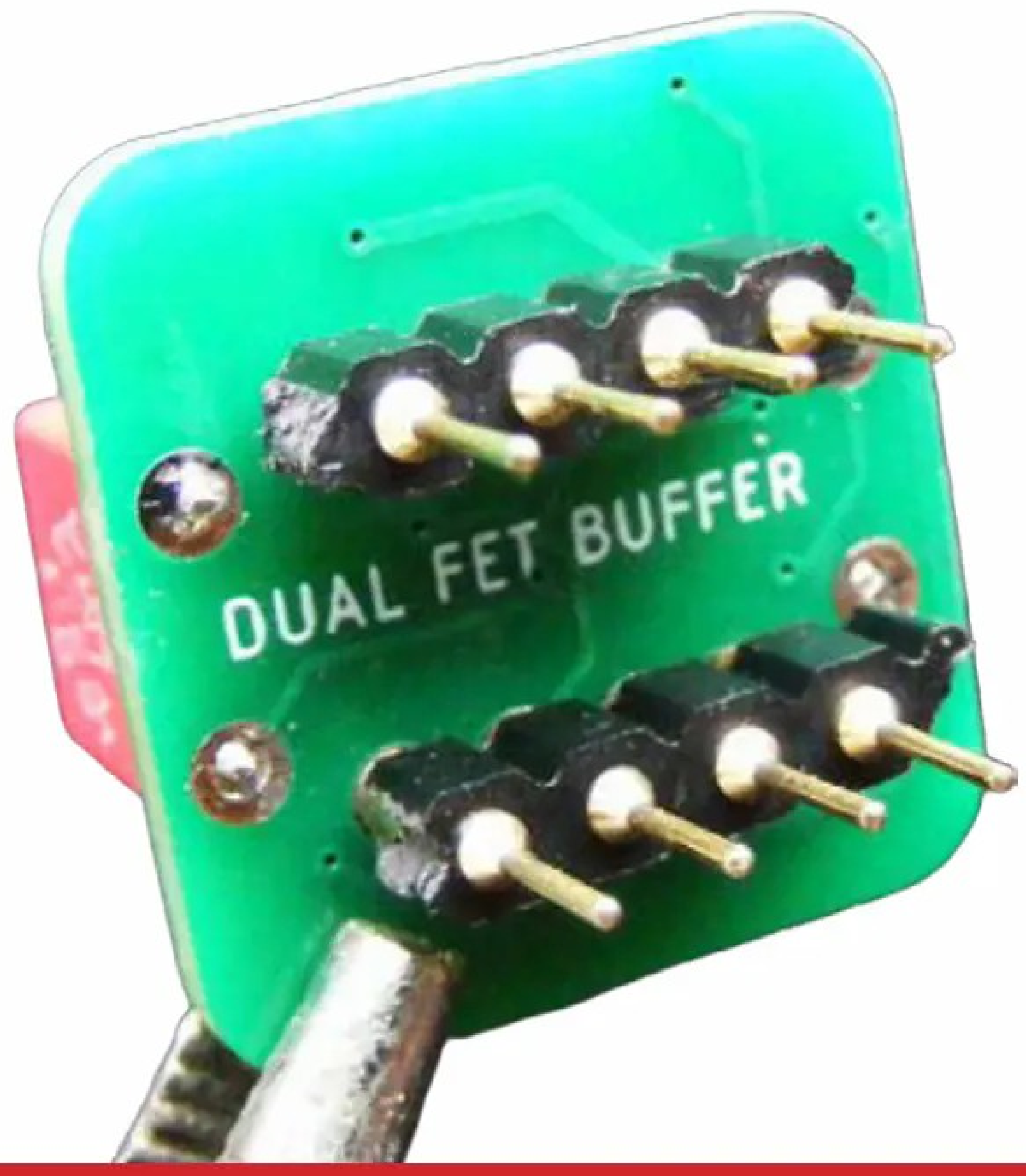
To test it, power it up and check that the supply current and output offset voltages are correct. If so, it should pass audio with symmetrical clipping (if the level is high enough). If there is a problem, it is normally due to incorrect resistor values having been used, or solder bridges or dry joints on the JFET pins.

High-spec Mosfet buffer

This relatively complex circuit, shown in Fig.6, will beat most expensive JFET audio op amps. Its operation was fully explained in the May 2024 issue. It's basically a JFET long-tailed pair, TR1/TR2, loaded by a current mirror (TR4/TR5) based on a PNP dual transistor, the HN1A01F. Fig.7 shows the distortion plot.

Each PNP transistor in the package is rated at 50V and 150mA, so it's similar to two BC556s, and it's cheap

Photo 3: it's essential to get the header strips straight so it fits in an IC socket.


DUAL FET BUFFER

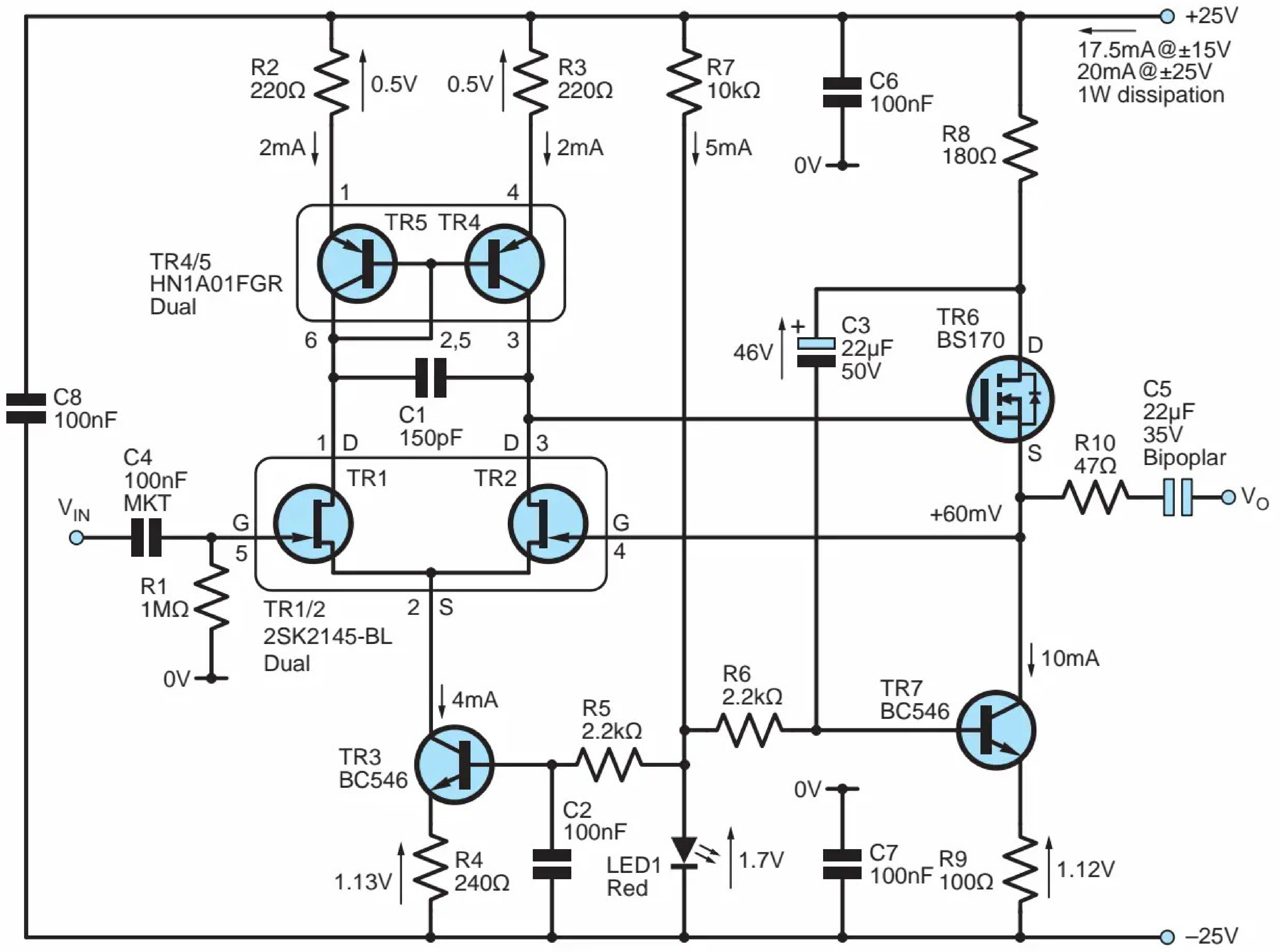

+25V
17.5mA@±15V
20mA@±25V
1W dissipation
R2
220Ω
0.5V
0.5V
R3
220Ω
R7
10kΩ
C6
100nF
2mA
2mA
5mA
0V
R8
180Ω
1
4
TR5 TR4
TR4/5
HN1A01FGR
Dual
6
2,5
3
46V
C3
22µF
50V
TR6
BS170
D
C8
100nF
C1
150pF
1 D
D 3
C5
22µF
35V
Bipoplar
C4
100nF
MKT
V_{IN}
TR1
TR2
G
5
G
4
S
R10
47Ω
+60mV
V_O
R1
1MΩ
0V
TR1/2
2SK2145-BL
Dual
2 S
10mA
4mA
R5
2.2kΩ
R6
2.2kΩ
TR7
BC546
TR3
BC546
0V
C2
100nF
1.13V
R4
240Ω
LED1
Red
1.7V
C7
100nF
R9
100Ω
1.12V
–25V

Fig.6: the Mosfet buffer provides very high performance from a more complex circuit.

Parts List – High-spec Mosfet buffer

1 double-sided 55 × 30mm PCB coded AO-OCT25-02
2 2-pin 0.1-inch (2.54mm) pitch polarised headers
1 3-pin 0.1-inch (2.54mm) pitch polarised header

Semiconductors
1 2SK2145-BL dual N-channel JFET (TR1/TR2) **[Mouser 757-2SK2145BLTE85LF]**
2 BC546 NPN small-signal transistors (TR3, TR7)
1 HN1A01F dual 50V 150mA PNP bipolar transistor, SM6 (TR4/TR5) **[Mouser 757-HN1A01F-GRTE85LF]**
1 BS170 60V 500mA N-channel Mosfet (TR6) **[Mouser 512-BS170]**
1 standard 5mm red LED (LED1)

Capacitors
1 22μF 50V radial electrolytic, 2.5mm pitch (C3)
1 22μF 35V bipolar electrolytic (C5)
4 100nF 50V X7R ceramic, 5mm pitch (C2, C6-C8)
1 100nF 63/100V ±10% polyester film (C4)
1 150pF ±5% C0G/NP0 ceramic, 5mm pitch (C1)

Resistors (all ¼W ±1% metal film axial)
1 1MΩ (R1)
1 10kΩ (R7)
2 2.2kΩ (R5, R6)
1 240Ω (R4)
2 220Ω (R2,3)
1 180Ω (R8)
1 100Ω (R9)
1 47Ω (R10)

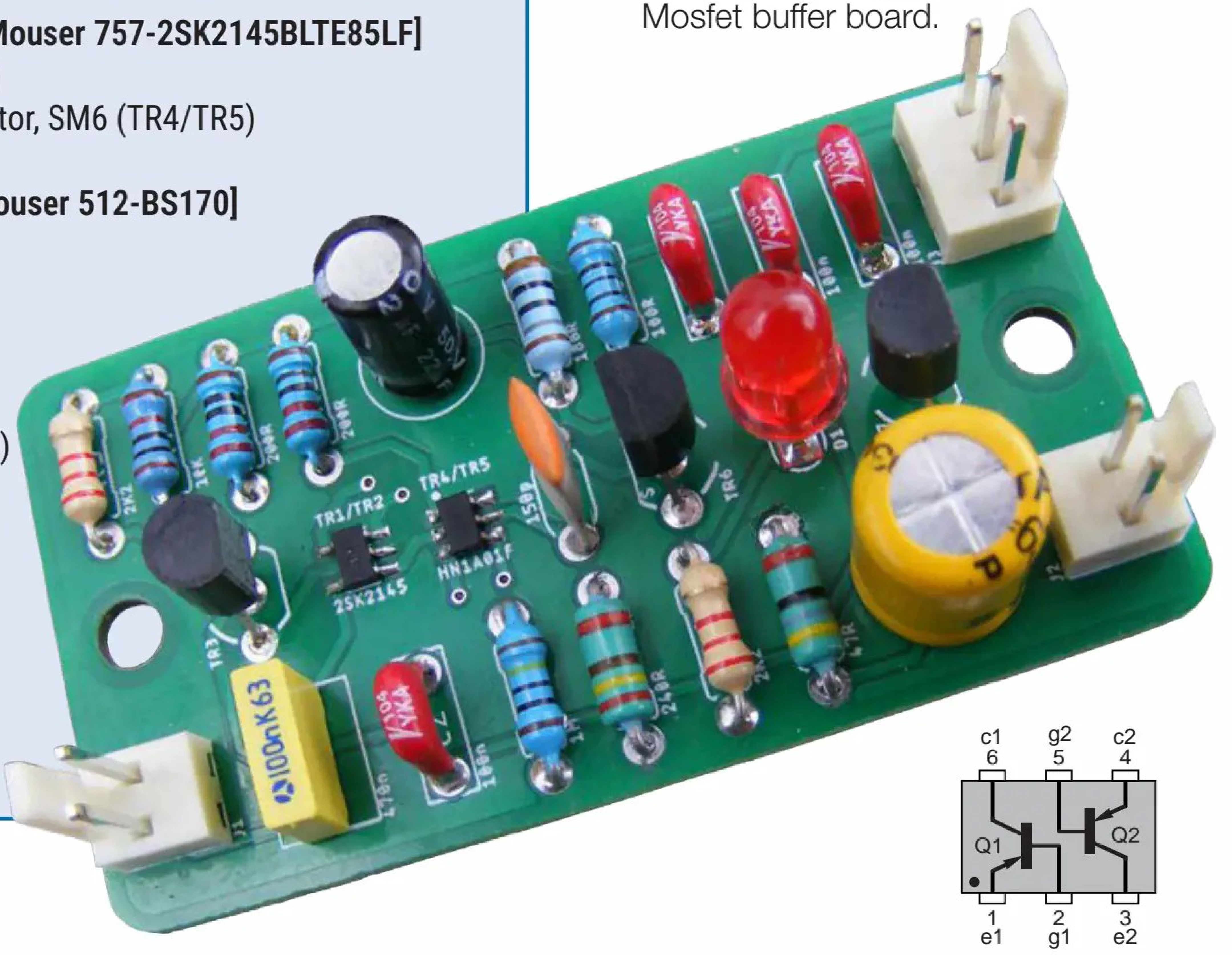


Photo 4: the single-channel Mosfet buffer board.

at 10p each in 100s. The long-tailed pair's non-inverting output then drives a Mosfet source-follower output stage (TR3) loaded by a modulated current sink (TR7). This is reasonably short-circuit-proof, but only on the negative cycle.

The output impedance is less than 1Ω, but a resistor is in series with the output to help with stability with long screened leads. If a low output impedance is needed to drive output transformers, the 47Ω output resistor (R10) can be bypassed with a small inductor of around 4μH.

If the unit is to be DC-coupled (as is also essential for audio transformer driving), output capacitor C5 should be omitted. The DC offset will then have to be trimmed out by tweaking the values of R3 and R4. I left out the preset that was in the original design, assuming AC coupling would be used in most cases.

Construction

The assembled board shown in Photo 4 is relatively simple, given the circuit, and the corresponding overlay is shown in Fig.8. The SMT transistors are the first components to solder; otherwise, protruding components will get in the way of your iron.

Bipolar transistor TR4/TR5 is unique in that the package can be mounted in two possible positions, with pin 1 up or down. It will work either way because the two transistors are connected such that the pinout is symmetrical (as shown in the diagram above).

Then fit the rest of the leaded components in the usual way, starting with the lowest-profile devices first.

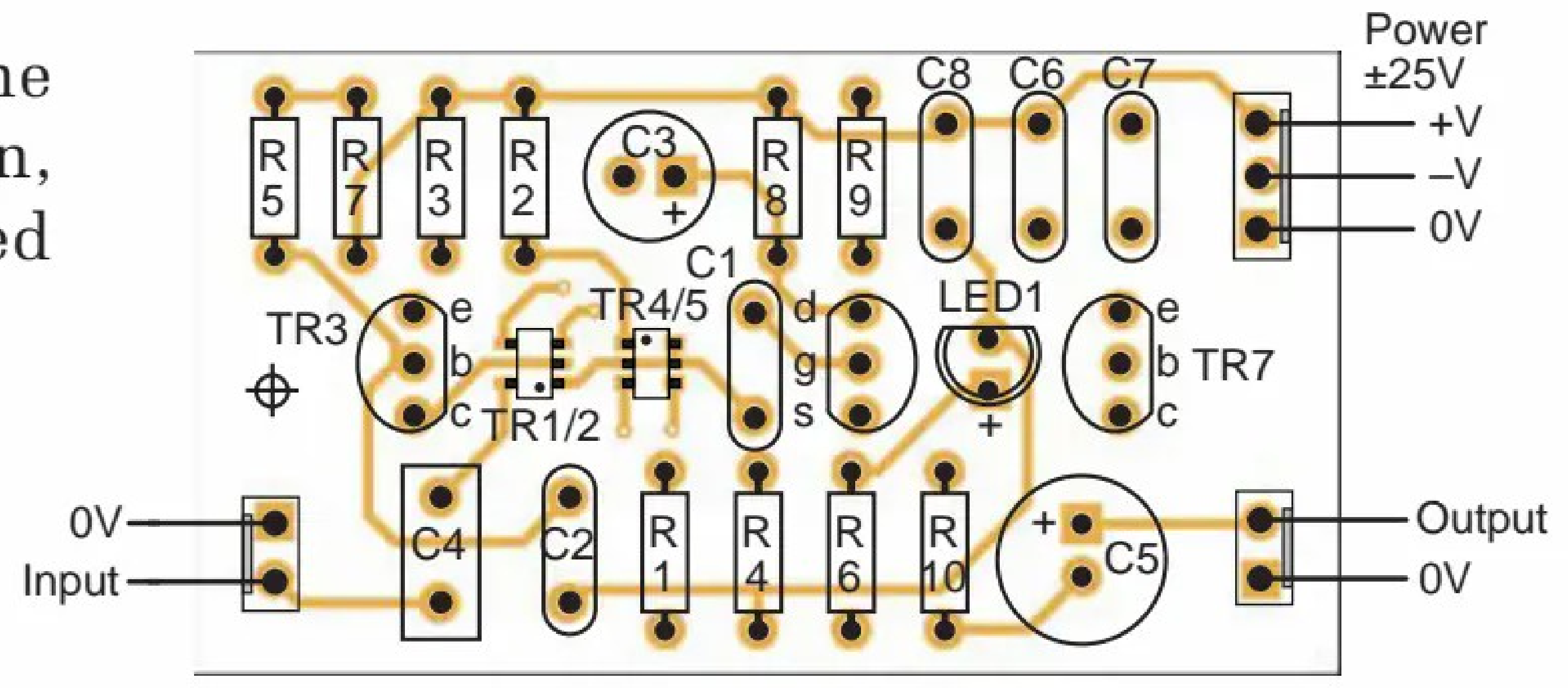


Fig.8: the high-spec Mosfet buffer PCB overlay.

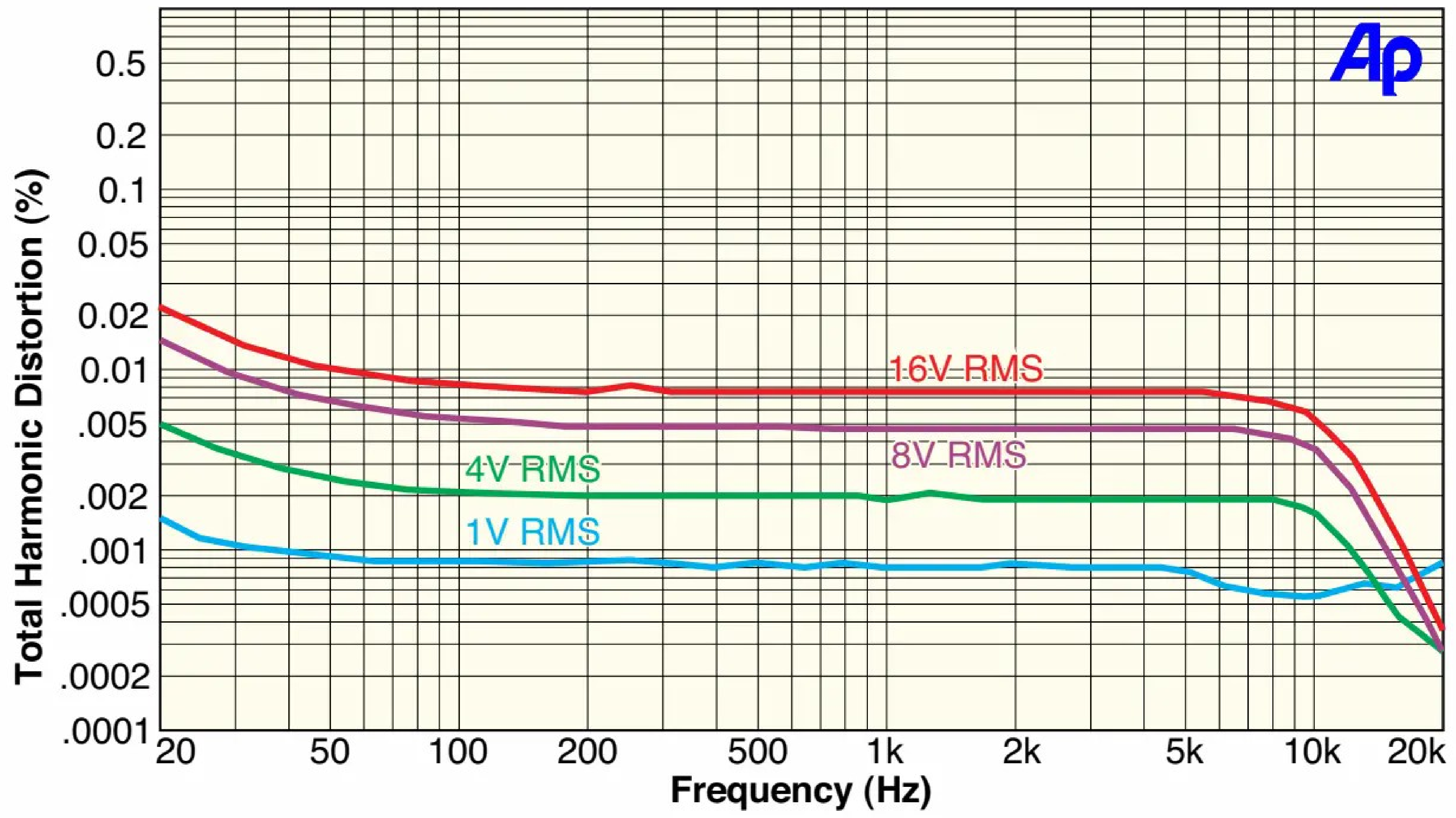


Fig.7: the distortion plots for the Mosfet buffer. The load is 600Ω for all curves; 16V RMS is just on the verge of clipping (45.4V peak-to-peak, delivering 430mW).

Testing

The board should be tested with a bench power supply with current limiting set around 300mA. Current consumption on both rails should be around 20mA, and the DC offset (measured on R10) should be less than ±100mV. If oscillation occurs, it can usually be fixed by inserting a gate-stopper resistor of 1kΩ in series with the input.

A stereo buffer

I've also had a stereo board designed based on a minimised version of the JFET input circuit in the April 2024 issue (see Fig.27 on page 63 of that issue), which is easier to build and use than two separate boards. The

Parts List – Dual Mosfet buffer

1 double-sided 65 × 52mm PCB coded AO-OCT25-03
2 2-pin 0.1-inch (2.54mm) pitch polarised headers
2 3-pin 0.1-inch (2.54mm) pitch polarised headers
1 4-pin 0.1-inch (2.54mm) pitch polarised header

Semiconductors
2 2SJ113 N-channel JFETs (TR1, TR4) **[AliExpress]**
2 BC556 PNP small-signal transistors (TR2, TR5)
2 BC546 NPN small-signal transistors (TR3, TR6)
2 standard 5mm red LEDs (LED1, LED2)

Capacitors
2 100µF 35V non-polar radial electrolytic, 5mm pitch (C4, C13)
2 100µF 25V radial electrolytic, 5mm pitch (C8, C7)
2 22µF 35V radial electrolytic or tantalum bead, 2.5mm pitch (C3, C12)
2 10µF 6.3V tantalum bead, 5mm pitch (C5, C14)
1 220nF X7R ceramic, 5mm pitch (C6)
2 100nF ±10% polyester film, 5mm pitch (C1, C10)
2 270pF ±5% NP0/C0G ceramic, 2.5mm pitch (C9, C15)
2 100pF ±5% NP0/C0G ceramic, 2.5mm pitch (C2, C11)

Resistors (all ¼W ±1% metal film axial)
2 1MΩ (R2, R13)
2 100kΩ (R11, R20)
2 6.8kΩ (R8, R19)
4 2.2kΩ (R3, R7, R14, R18)
2 620Ω (R1, R12)
2 300Ω (R5, R16)
2 220Ω (R9, R10)
2 100Ω (R4, R15)
2 33Ω (R6, R17)

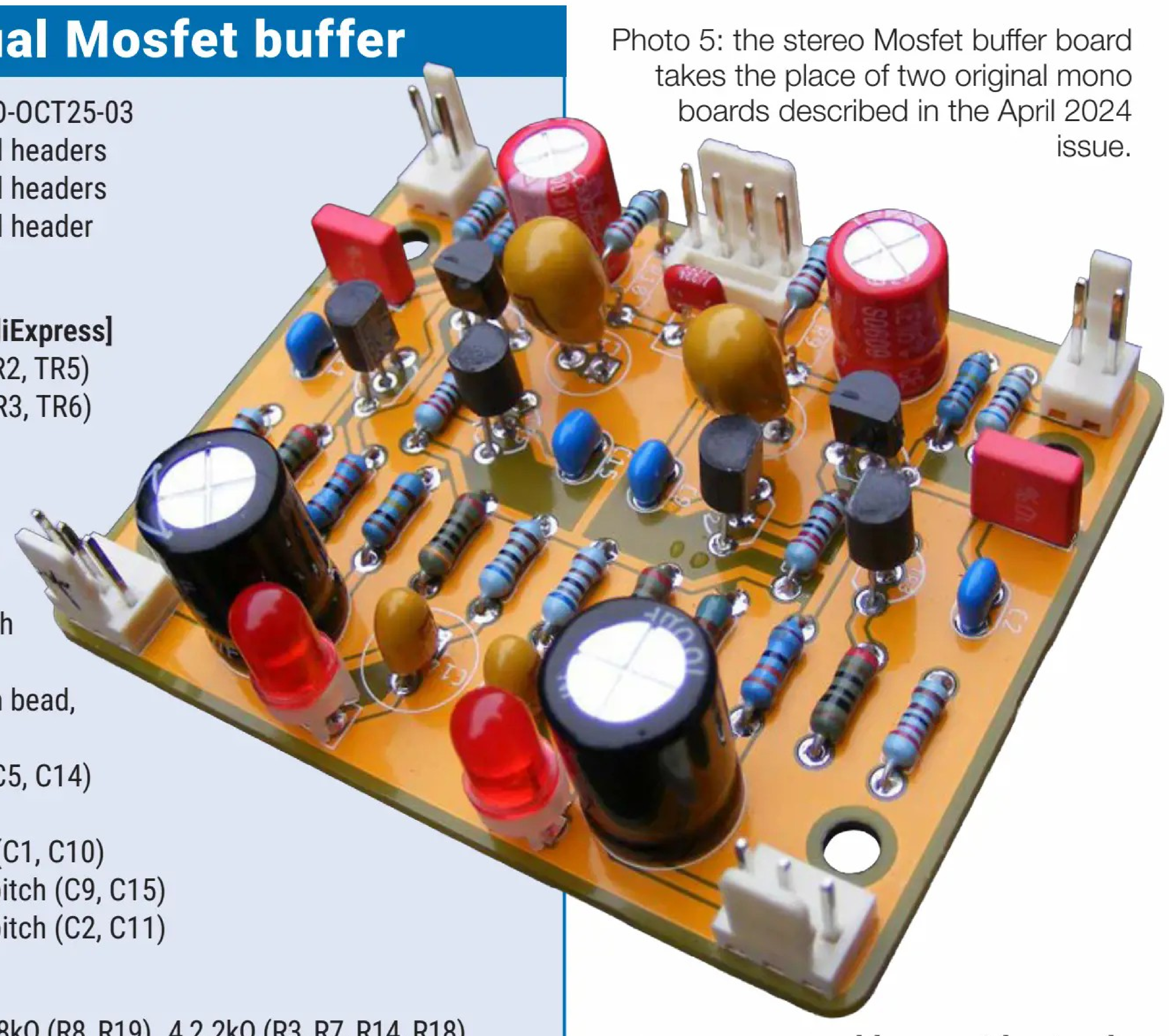


Photo 5: the stereo Mosfet buffer board takes the place of two original mono boards described in the April 2024 issue.

circuit is repeated in Fig.9 because the component numbering is different.

I thought it could be simplified further by employing one voltage reference for the current sinks. This caused problems in that if one side clipped, it transferred a glitch to the other side. I've noticed this problem on a few dual op amp chips too.

The overlay is given in Fig.10, while Photo 5 shows the completed board. There is nothing untoward about the construction, and it should appeal to those who love symmetrical PCBs.

Buffer board applications

A popular item in minimalist Hi-Fi and studio monitoring setups is simply a stereo volume control in a box, sometimes called a passive preamp. The potentiometer often used is the Alps Blue Velvet (shown in Photo 6) on account of its smooth audio taper and channel matching at low levels.

There is a catch, however, in that it is usually only available in 50kΩ. This results in an output impedance varying from 25kΩ to 4.5kΩ as it is rotated. This can interact with long cables and the input impedance of the power amp or active speaker being driven, causing high-frequency loss, pot-law deviation and distortion. These are common problems with simple passive preamps.

The solution, of course, is to buffer the output of the pot. This now makes it an active preamp, which has the problem of needing a power supply. *C'est la vie.*

Sourcing the parts

All three PCBs and the parts that mount on them are available from the AOShop (see page 13).

These PCBs are also available at the same price from our PCB Service, in case you want to order them along with some for other projects.

Coming up next month

Enough of clean, low-distortion

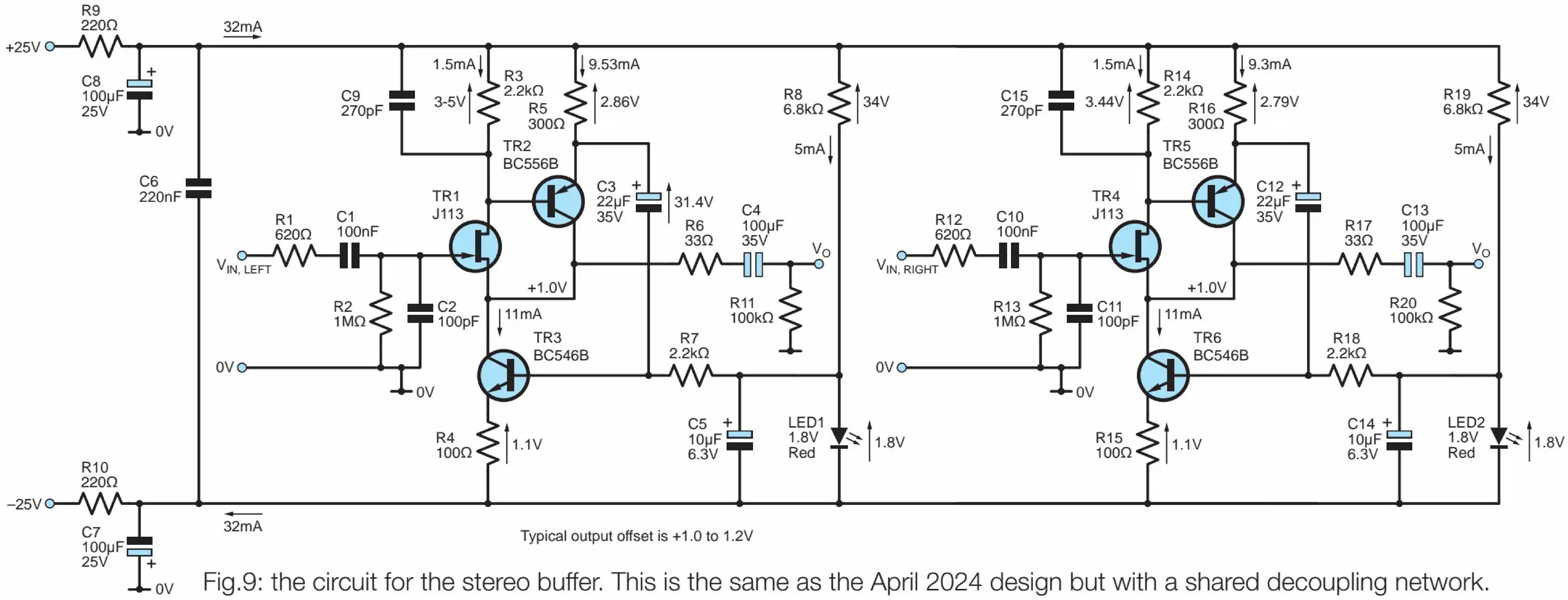

+25V
R9
220Ω
32mA
C8
100µF
25V
0V
C6
220nF
C9
270pF
1.5mA
3-5V
R3
2.2kΩ
9.53mA
R5
300Ω
2.86V
TR2
BC556B
TR1
J113
R1
620Ω
C1
100nF
V_IN, LEFT
R2
1MΩ
C2
100pF
0V
0V
C3
22µF
35V
31.4V
R6
33Ω
C4
100µF
35V
V_O
+1.0V
11mA
TR3
BC546B
R7
2.2kΩ
R11
100kΩ
R4
100Ω
1.1V
C5
10µF
6.3V
LED1
1.8V
Red
1.8V
R8
6.8kΩ
34V
5mA
−25V
R10
220Ω
32mA
C7
100µF
25V
0V
Typical output offset is +1.0 to 1.2V
C15
270pF
1.5mA
3.44V
R14
2.2kΩ
9.3mA
R16
300Ω
2.79V
TR5
BC556B
TR4
J113
R12
620Ω
C10
100nF
V_IN, RIGHT
R13
1MΩ
C11
100pF
0V
0V
C12
22µF
35V
R17
33Ω
C13
100µF
35V
V_O
+1.0V
11mA
TR6
BC546B
R18
2.2kΩ
R20
100kΩ
R15
100Ω
1.1V
C14
10µF
6.3V
LED2
1.8V
Red
1.8V
R19
6.8kΩ
34V
5mA

Fig.9: the circuit for the stereo buffer. This is the same as the April 2024 design but with a shared decoupling network.

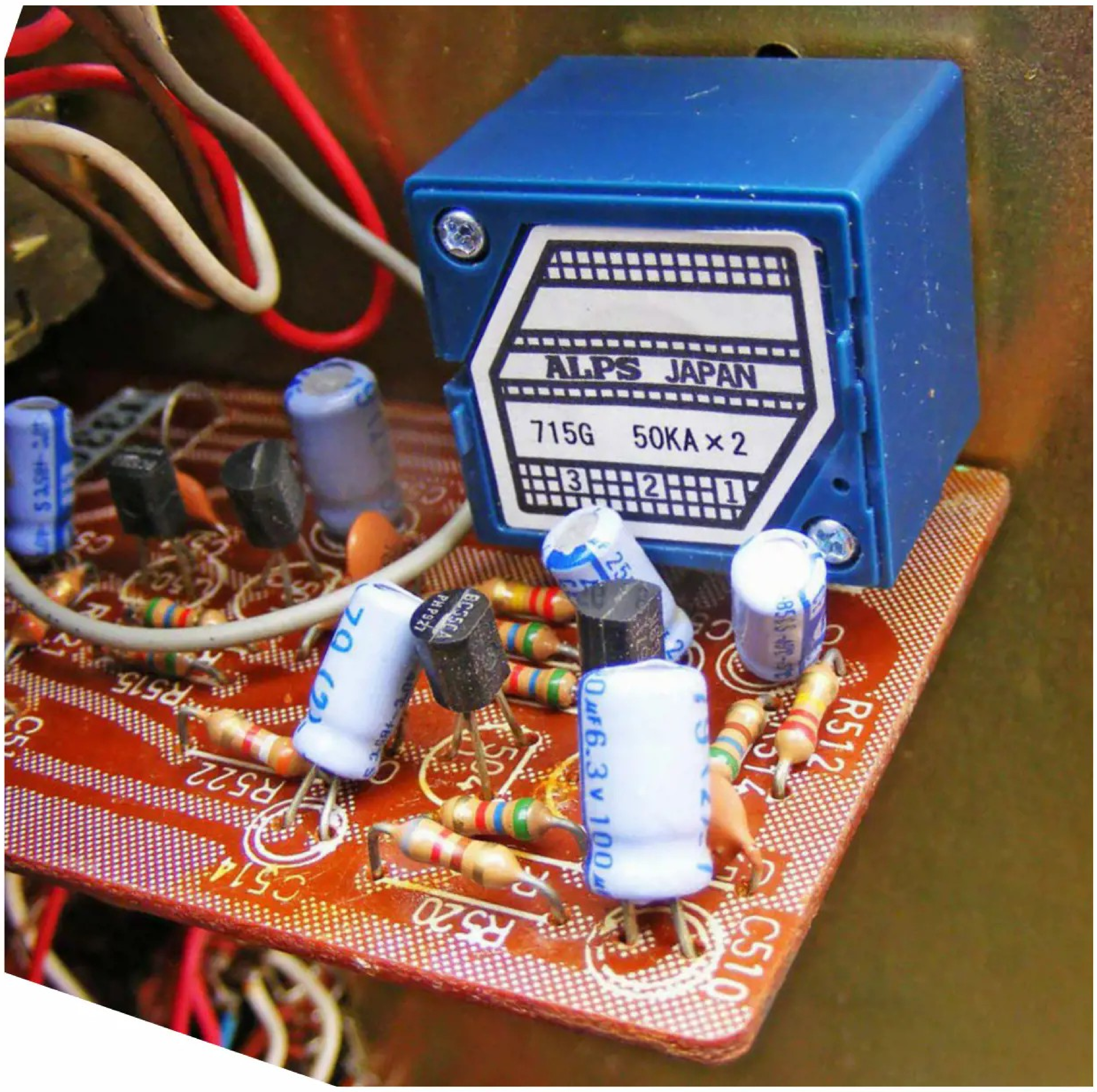

ALPS JAPAN
715G 50KA × 2
3 2 1
R512
R515
R522
R520
C514
C510

Photo 6: an Alps volume pot installed in a NAD 3030 amplifier. It needs a low-impedance drive and high-impedance load to minimise the effect of track resistance tolerance. These pots' rotational characteristics are optimised when they are fully buffered.

living for now. In the following article, we'll 'get dirty' with a transformer-coupled germanium transistor fuzz box!

Fuzz boxes are used by musicians, especially in rock and jazz music, to alter the sound of instruments (usually an electric guitar or bass) to add harmonics. These give the sound more of a 'bite'.

It is especially helpful for a bass guitar because the higher-frequency harmonics can be more audible in the mix, while the lower frequencies are often more felt than heard. Germanium fuzz boxes are generally thought to give the best sound. **PE**

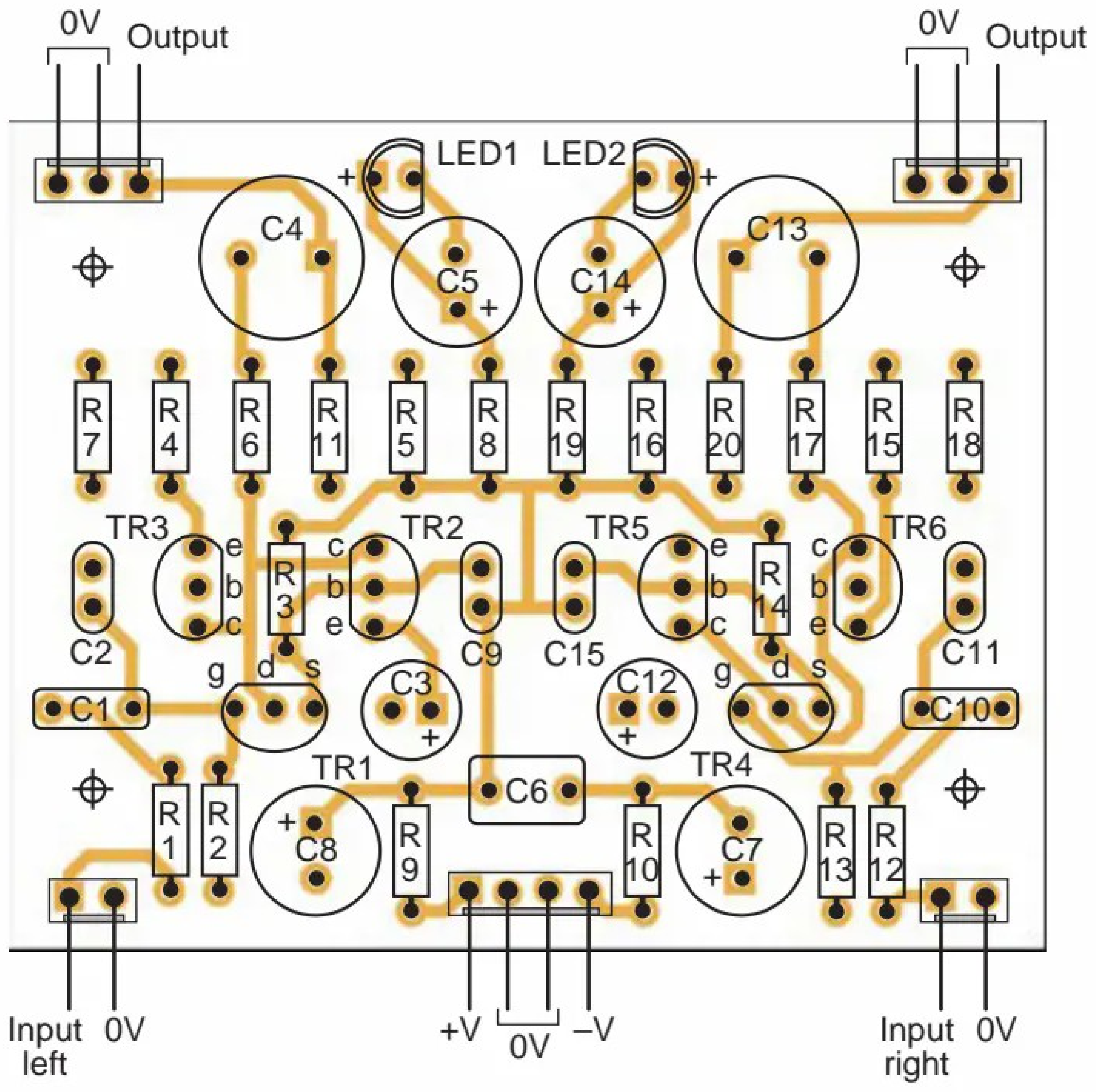
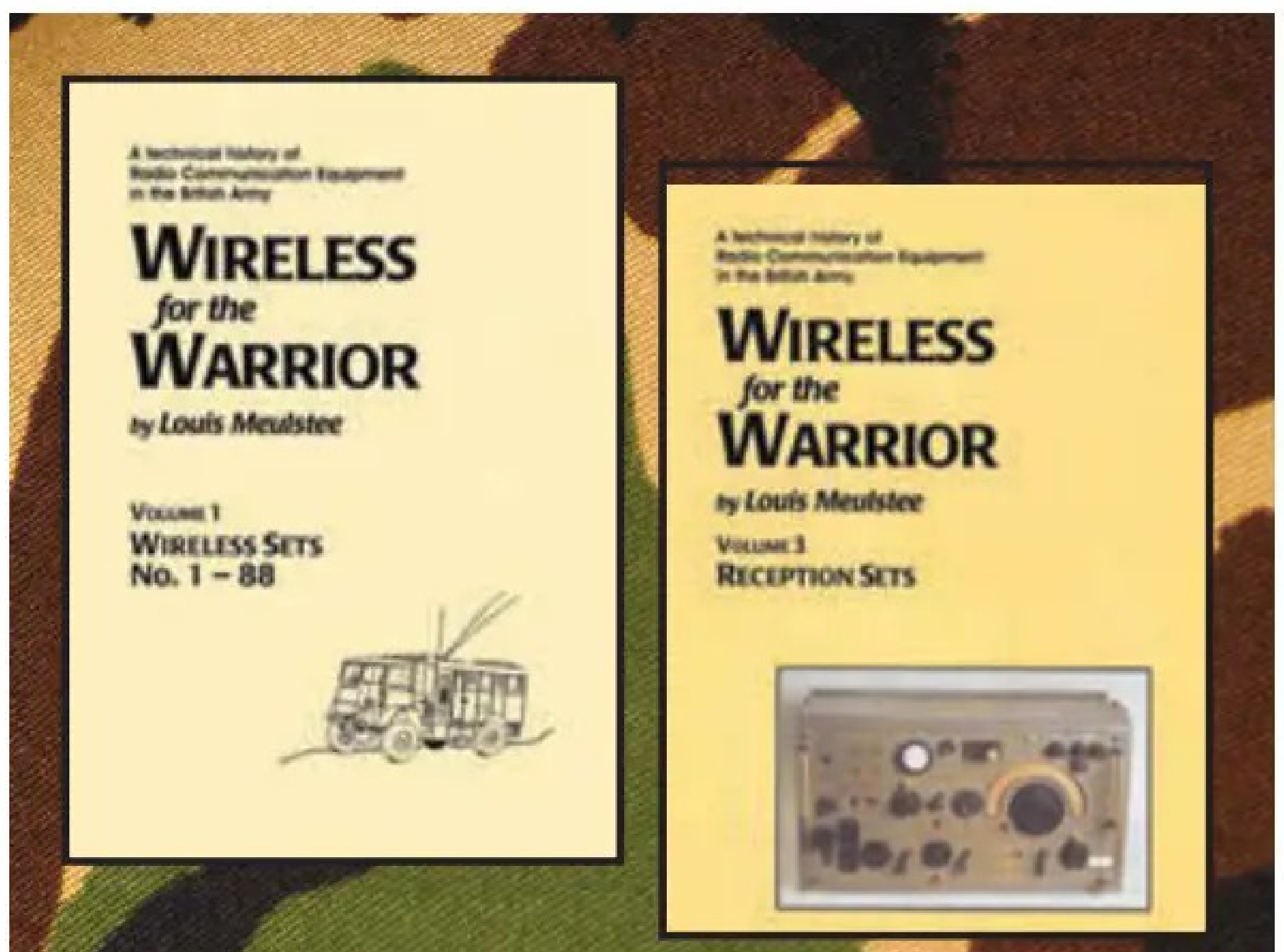

0V
Output
0V
Output
LED1
LED2
C4
C5
C14
C13
R7
R4
R6
R11
R5
R8
R19
R16
R20
R17
R15
R18
TR3
TR2
TR5
TR6
e
b
c
R3
R14
C2
C9
C15
C11
C1
C3
C12
C10
g
d
s
TR1
TR4
C6
R1
R2
C8
R9
R10
C7
R13
R12
+
Input left
0V
+V
0V
–V
Input right
0V

Fig.10: the PCB overlay for the stereo version of the buffer.


A technical history of Radio Communication Equipment in the British Army
WIRELESS for the WARRIOR
by Louis Meulstee
Volume 1
Wireless Sets No. 1 – 88
Volume 3
Reception Sets

The Wireless for the Warrior books are references for the history and development of radio communication equipment used by the British Army from the very early days of wireless up to the 1960s.

Volume 1 covers early transmitters and transceivers used between 1932 & 1948. Volume 3 covers army receivers from 1932 to the late 1960s. The book not only describes receivers specifically designed for the British Army, but also the Royal Navy and RAF.

Volumes 1 & 3 are still available, but stocks are running out, so they won't last long. Order a printed copy now from:

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By Andrew Levido

Precision Electronics

Part 9: System Design

In this last article in the Precision Electronics series, we look at the design of a precision electronics system from the big-picture perspective. We have already covered a lot of the building blocks; this will bring them together and show how to approach the design of a whole system.

To do this, we will look at a practical example of moving from a high-level specification to developing design goals for each circuit block, then dive in detail into a couple of the blocks for good measure.

Before we get into it, I want to summarise some of the key tips and tricks we have learned from previous articles that might help guide our design. These are not hard-and-fast rules; they are just things I have found it helpful to keep in mind:

- Precision and accuracy are different things. Precision is all about errors and repeatability, while accuracy is all about calibration and traceability.
- Break circuits down into bite-sized chunks to perform error analysis. Things can get overwhelming if the circuit being analysed is too complex.
- When like quantities with errors add or subtract, add the absolute errors. When quantities with errors multiply or divide (like an input offset through a gain), add the relative (proportional) errors.
- Completely uncorrelated random errors (like white noise) can be added as the root sum of squares – either absolute or relative, depending on the application. This also applies to calculating a DAC's or ADC's total unadjusted error (TUE) figure.
- You can easily calibrate or trim out fixed errors like offset and gain errors. It is harder to do so for non-linearities or errors that change with temperature. The highest-precision devices use this technique extensively, often performing calibration before every reading.
- In general, keep the span of precision signals away from the power rails or the ends of ADC and DAC ranges where they coincide with the supply rails. This is one place where non-linear errors love to hide.
- Read data sheets carefully, including the graphs. Manufacturers don't usually highlight the shortcomings of their parts in the headline specs. Use worst-case errors (not typical values) unless you have a good reason not to.
- Use larger signals and lower gains where you have a choice. Input-side errors are magnified by gain stages. The signal-to-noise ratio can be improved by using larger signals.
- Reduce noise by limiting circuit bandwidth and using lower-value resistors where possible. Oversampling and averaging is a useful form of bandwidth limitation.
- Reduce cost by using components with no higher precision than you absolutely need. Use an error budget spreadsheet to understand the major contributors to error and focus on these first.

The process of top-down design is kind of the opposite of what we have

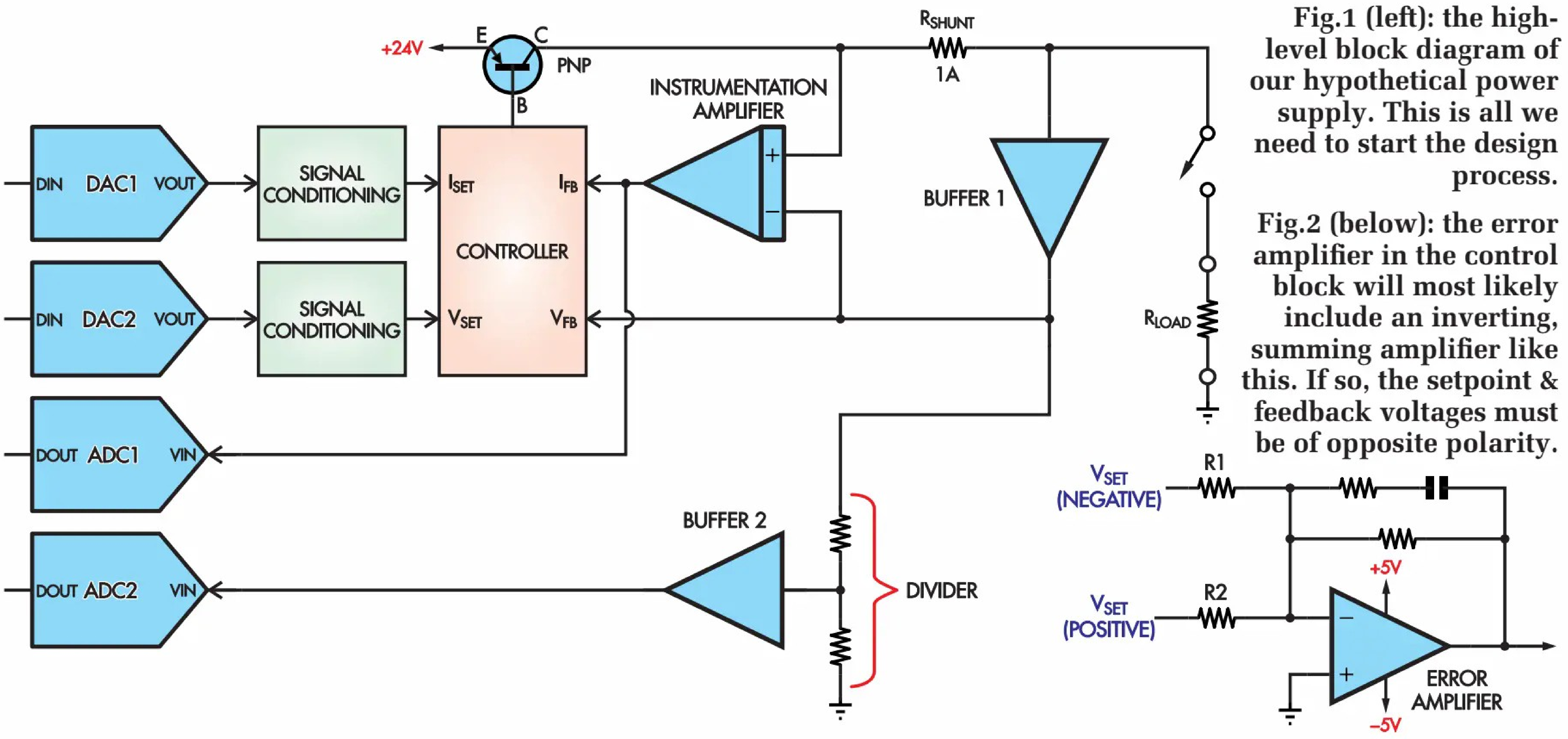


Fig.1 (left): the high-level block diagram of our hypothetical power supply. This is all we need to start the design process.

Fig.2 (below): the error amplifier in the control block will most likely include an inverting, summing amplifier like this. If so, the setpoint & feedback voltages must be of opposite polarity.

done so far, where we looked at individual circuits and calculated their errors. This time, we will divide the overall system error up to develop an error 'budget' that will guide the design of each subsystem.

This sounds very straightforward, but in practice there is almost always a certain amount of iteration (even to the point sometimes of revising the top-level specifications).

A simple power supply

The example I want to work with is a simple power supply that can source a voltage in the range of 0-20V with a maximum current of 1A. We are not going to fully design this power supply here – we are just aiming to develop the error budget and look at the system considerations including coming up with a calibration strategy.

I will go through a more detailed component selection and analysis for a small part of the circuit, just to show how I go about it.

Let's start with some target specifications. We want to be able to set the voltage in 100mV steps and the current limit in 10mA steps. This corresponds to a modest precision of ±0.5% for the voltage and ±1% for the current setpoint. However, we want to be able to measure the voltage and current to a precision of ±0.1% (±20mV and ±1mA respectively).

These specs are all defined at nominal temperature (25°C). For simplicity's sake, we will assume that they should be no worse than 150% of nominal over the operating temperature range. This is a laboratory instrument, so I am going to arbitrarily decide that a range of 5°C to 45°C (25°C ±20°C) will be adequate.

High-level circuit design

Designs should start with a simplified block diagram, as shown in Fig.1. Here, a microcontroller (not shown) feeds a pair of DACs which, together with the signal conditioning blocks, create the voltage and current setpoints that are applied to the power supply control circuit. This circuit regulates the output voltage and manages the current limiting.

Feedback of the output voltage and current is necessary to allow the control circuit to regulate properly.

The voltage feedback for the control circuit comes from the power supply output via high-impedance buffer 1. This unity-gain buffer is required to minimise the current drawn from the output terminal, as this will affect the current measurement. The same buffer provides the voltage measurement input to ADC2 via a voltage divider and a second unity-gain buffer.

Current sensing is provided via a high-side shunt resistor and instrumentation amplifier. This feeds both the control circuit and the current measurement function via ADC1. We will need two DACs since we have to provide both setpoints simultaneously, but we can get away with one ADC with an input multiplexer, since we can sample the current and voltage alternately.

Error amplifier

The control circuit is most likely going to contain error amplifiers made using an inverting, summing amplifier like that shown in Fig.2. This topology has two implications for our design. First, it means that the senses of the setpoint and feedback signals need to be opposite, so the error amplifier sees the difference between the output and the setpoint.

If the setpoint is negative, as shown, a setpoint increase (ie, a more negative voltage) will cause the error amplifier output to increase in a positive direction, driving the output

Two-point calibration

The figure below shows a system like the voltage setpoint and regulation circuit in our example power supply. The ideal transfer function for the system is y = kw, where k is the design gain (for example, some number of millivolts per LSB), w is the input digital code and y is the output quantity. There is gain, but no offset, in this ideal system.

A real system may have both a gain error and an offset error, as shown on the bottom of the figure. Here, the gain (m) is close but not equal to the ideal gain, k, and there is a non-zero offset, b. The firmware correction block shown on the left takes in the input code w and applies correction gain n and offset c to produce a corrected code x, such that the output of the hardware system y has the ideal relationship with the input code w.

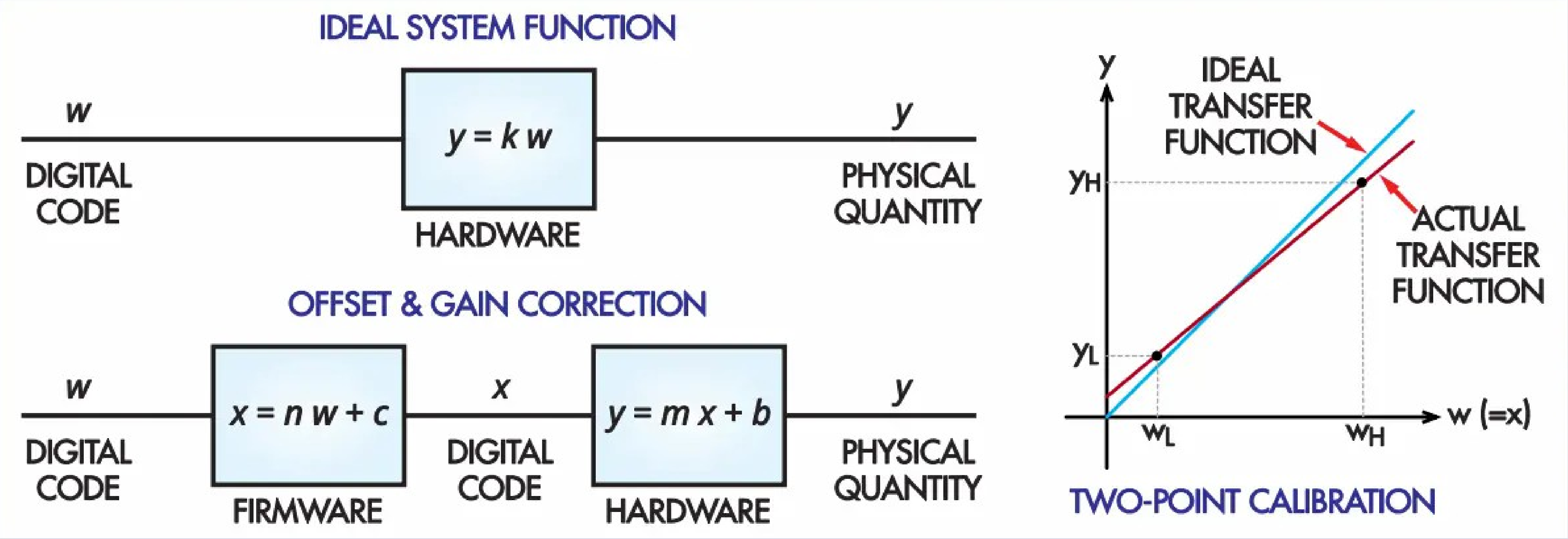
If we measure the hardware system's transfer function (ie, find the gain [m] and offset [b]), we can calculate the compensating gain n and offset c. Two-point calibration is the simplest process that allows us to calculate these correction factors.

The calibration cycle starts by setting the correction gain n to unity and the correction offset c to zero. This means DAC code x will equal the raw code w. The calibration firmware then sets the code to a value near the bottom end of the span (w_L), as shown in the figure, and we measure the output quantity (y_L) using an external meter. This value is provided to the firmware.

Next, the calibration routine sets code w to one near the top of the span (w_H). This is also measured and entered into the instrument.

Now the firmware can calculate the hardware's transfer function coefficients using the equations $m = (y_H - y_L) \div (x_H - x_L)$ and $b = y_L - mx_L$. With this done, the correction factors can be calculated from the relations $n = k \div m$ and $c = -b \div m$. These factors would be stored in non-volatile memory and used to correct the DAC code to compensate for the gain and offset errors in the hardware.

Two-point calibration has the advantage of being very simple to implement. You can improve on it by using more points (called multi-point calibration), and deriving the hardware transfer function from a suitable line-of-best-fit algorithm.



voltage or current limit higher. Similarly, if the feedback level increases, the error amp output will be forced lower, reducing the output voltage or current limit.

Secondly, the error amplifier summing junction (the op amp's inverting input) sits at 0V and actually sums currents determined by the input voltages and the values of the two input resistors, R1 and R2.

The result is that the full-scale setpoint voltage and full-scale feedback voltages do not have to be the same, since we can adjust the values of the resistors to make their full-scale currents equal.

We do have to be concerned with the precision of the ratio between the resistor values, as any deviation will cause an error in the setpoint. Fortunately, we do not have to worry about anything further into the controller circuity, as any offset and gain errors here are eliminated by the action of the feedback.

The same is true for non-linearities in this part of the circuit, although if they are extreme, they can impact control loop stability.

Block diagrams & error budget

Fig.3 shows the circuit in Fig.1 broken down into four 'signal chains', one each for the voltage and current controls, and one each for voltage and current sensing. This is a great way to get a handle on just which circuit block impacts the overall device precision.

Some of the blocks appear in multiple chains; for example, the output voltage buffer appears in all the chains. In other cases, like the setpoint signal conditioning block, there are separate but identical blocks.

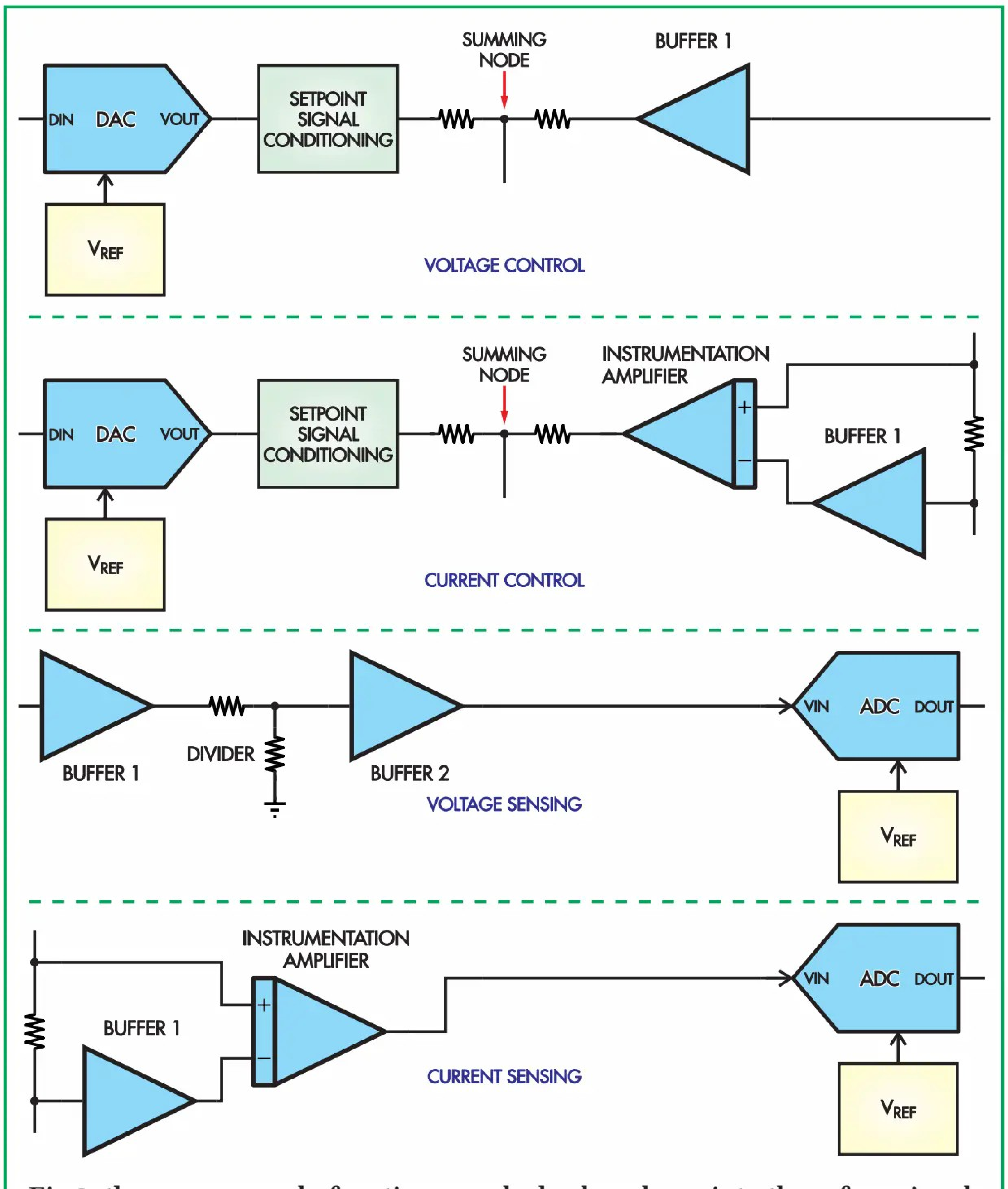


Fig.3: the power supply functions can be broken down into these four signal chains. Some of the circuit blocks appear in multiple chains, complicating the process of assigning error budgets to each one.

The task, then, is to take the overall precision specification for each signal chain and allocate it to the various circuit blocks. The duplication and repetition of blocks described above makes this a slightly complex process.

I used an error budget spreadsheet like that of Table 1. I have listed the unique circuit blocks down the left-hand side and the four signal chains across the top.

Where a particular circuit block appears in a signal chain, I entered the appropriate number in the matrix. I then made a stab at setting an error budget for each circuit block down the left side, while checking the totals in blue along the bottom compared to the targets shown in purple.

It is best to start from the most difficult chain (the current sensing chain in this case) and work from there. You can see that some of the error budgets end up better than the target, which is OK at this stage.

There is no point in allocating the error budget down to the nth degree in this process. You can see that the budget here shows the current-sensing signal chain to be just over the target. This is near enough for this stage in the design cycle.

It does take a bit of experience to work out what are reasonable error assumptions for each particular block, but hopefully, some of the previous articles in this series have given you a feel for it. Clearly, a unity-gain buffer using a zero-drift op amp will have a much lower error than a 16-bit ADC.

The next step is to work through the design, block by block, selecting components and topologies that will meet the targets for each subcircuit. I recommend starting with the subsection where you think it will be most difficult to achieve the target precision. This may not be the area with the highest precision requirement.

Calibration strategy

The errors in Table 1 are trimmed errors, so we need to have some idea of our calibration strategy to translate these to untrimmed errors for the design process.

To keep things simple, for this power supply, I plan to perform calibration manually when we build the supply and every now and again thereafter. This eliminates the need for in-system calibration circuitry.

Part 9 – overall system design

However, this implies that our calibration will only be able to improve fixed gain and offset errors at the nominal temperature.

The minimum required to do this is two-point calibration (described in detail in the accompanying panel). This means calculating and storing an offset and a gain calibration value for each of the four signal chains.

In practice, the calibration of the setpoint and measurement chains can be done simultaneously. For example, to calibrate the two voltage signal chains, we would set the voltage at some low value (say 1.0V) and measure the actual terminal voltage with an external meter with sufficient resolution. We would do the same at some high value (say, 19.0V).

Once entered into the system, these two values can be used to calculate calibration coefficients for the setpoint chain by comparing them with the nominal setpoint, and the measurement chain by comparing them with the values read by the ADC.

Given this calibration strategy, I have made the assumption that we can calibrate fixed offset and gain errors down to 5% of their untrimmed values (ie, we can calibrate out 95% of the error). Non-linear errors and temperature dependent errors are not reduced at all by this method.

Voltage setpoint design

I will go through the process in some detail for a small part of the circuit to show you the idea, starting with the voltage setpoint circuit, although I have included the full set of error calculations in Table 2.

Table 1 – error targets for each circuit block (in purple) and the resulting error (in blue)

Circuit Block	Error @ 25°C	Additional Error ±20°C	Voltage Setpoint	Current Setpoint	Voltage Sensing	Current Sensing
DAC	0.250%	0.125%	1	1		
Setpoint signal conditioning	0.100%	0.050%	1	1		
Summing node	0.013%	0.006%	1	1		
Buffer	0.013%	0.006%	1	1	2	1
Current sensing	0.040%	0.020%		1		1
Voltage divider	0.013%	0.006%			1	
ADC	0.050%	0.025%			1	1
Voltage reference	0.003%	0.001%	1	1	1	1
		Target error @ 25°C	0.500%	1.000%	0.100%	0.100%
		Target additional error ± 20°C	0.250%	0.500%	0.050%	0.050%
		Total @ 25°C	0.378%	0.418%	0.090%	0.105%
		Total additional ± 20°C	0.189%	0.209%	0.045%	0.053%
		Total error 5°C to 45°C	0.566%	0.626%	0.135%	0.158%

To do this, we need to define some full-scale voltage levels. We will start by assuming that we have a 3.3V logic supply for the microcontroller and ±5V analog supplies, as well as the 24V unregulated supply used for the output.

We would like to use a low-cost serial DAC, since the setpoint precision requirements are not strenuous. This will have to run from 3.3V to interface with the micro, so we will be limited to a full-scale analog range somewhere below this level. If we select a full-scale voltage of 2.5V, we will have a wide choice of low-cost external voltage references.

One of our precision design rules of thumb is that we should not trust analog values as they approach the power supply limits. We will have a DAC output that varies between 0V and 2.5V, depending on the input code.

Given a resistor-string DAC with a 3.3V supply, we can be reasonably comfortable at the top of the range, since there is plenty of headroom between the highest tap voltage (very close to 2.5V) and the analog supply rail (3.3V).

However, the bottom end of the DAC's resistor string will be grounded, as will the DAC's analog circuitry, so we cannot count on the lower end of the range. This is a problem because it suggests that we will lose accuracy near the bottom of the range and/or might not be able to set the output voltage or current limit right down to zero.

		At Nominal 25°C		Additional error over 25 ±20°C	
	Error	**Abs. Error**	**Rel. Error**	**Abs. Error**	**Rel. Error**
1	**DAC: MCP48FVB22, 12-bit, Two Channels, SPI**				
2	DAC Offset Error: ±15mV, 10µV/°C	15mV	0.750%	200µV	0.010%
3	DAC Gain Error: ±1%, 3ppm/°C	20mV	1.000%	60µV	0.003%
4	DAC INL: ±4LSB	2mV	0.098%	0mV	0.000%
5	**Trimmed Error: 5% of (Line 2 + Line 3, root sum squares) + Line 4**	**3.2mV**	**0.160%**		
6	**Temperature Drift Error: Line 2 + Line 3, root sum of squares**			**208.8µV**	**0.010%**
	Setpoint Signal Conditioning: TPA1834 Op Amp, Quad Zero Drift				
1	Op Amp Offset Error: ±7µV, ±0.04µV/°C	7µV	0.000%	800nV	0.000%
2	Op Amp Gain Resistor R1/R2: Vishay ACASA, 0.1%, 0.05% matched, 15ppm/°C	1mV	0.050%	600µV	0.030%
3	**Trimmed error 5% of (Line 1 + Line 2)**	**50.4µV**	**0.003%**		
4	**Temperature Drift Error: Line 1 × Line 2**			**600.8µV**	**0.030%**

Table 2 – detailed error calculations each signal block (see overleaf for the rest of the table; LSB = least significant bit). Absolute error values are with respect to 2V out

	Table 2 continued...	**At Nominal 25°C**		**Additional error over 25 ±20°C)**	
	Error	**Abs. Error**	**Rel. Error**	**Abs. Error**	**Rel. Error**
	Summing Node: RN73C2A, 0.1%, 10ppm/°C				
1	Summing Node Gain Error		0.200%		0.020%
2	**Trimmed error 5% of Line 1**		**0.010%**		
3	**Temperature Drift Error (Line 1)**				**0.020%**
	Buffer: TPA1834 Op Amp, Quad Zero Drift				
1	Op Amp Offset Error: ±7µV, ±0.04µV/°C	7µV	0.000%	800nV	0.000%
2	**Trimmed error 5% of Line 1**		**0.000%**		
3	**Temperature Drift Error (Line 1)**				**0.000%**
	ADC: ADS1115, ΔΣ, 16-bit, 4CH, I^2C				
1	ADC Offset Error: ±3LSB, 0.005LSB/°C	114.4µV	0.005%	3.1µV	0.000%
2	ADC Gain Error: ±0.15%, 40ppm°C	3.8mv	0.150%	2mV	0.080%
3	ADC INL: ±1LSB	38.1µV	0.002%		
4	**Trimmed error 5% of (Line 1 + Line 2, root sum squares) + Line 3**	**225.7µV**	**0.056%**		
5	**Temperature Drift Error (Line 1 + Line 2, root sum of squares)**			**2mV**	**0.080%**
	Current Sense				
1	Shunt Error 1Ω: VMP-1R00-1.0-U, 1%, 20ppm/°C	10mΩ	1.000%	400µΩ	0.040%
2	In Amp: INA821: V_{OS} ±35µv, 0.4V/°C	35µV	0.002%	8µV	0.002%
3	In Amp Input Voltage Error Total: Line 1 + Line 2	10mV	1.004%	408µV	0.041%
4	In Amp Gain Resistor R_G: ERA-6ARB333V (0.1%, 10ppm/°C)		0.100%		0.020%
5	In Amp gain error (0.015% ±35ppm/°C)		0.015%		0.070%
6	Total In Amp gain error (Line 4 × Line 5)		0.115%		0.090%
7	**Trimmed error 3% of (Line 3 × Line 6)**	**677.9µV**	**0.034%**		
8	**Untrimmable temperature drift error (Line 2 × Line 3)**			**2.6mV**	**0.131%**
	Voltage Sense Divider: RN73C2A, 0.1%, 10ppm/°C				
1	Divider gain error		0.200%		0.020%
2	**Trimmed error 5% of Line 1**		**0.010%**		
3	**Temperature drift error (Line 1)**				**0.020%**
	Reference: REF3425TD, 2.5V, 0.05%, 6ppm/°C				
1	V_{REF} error		0.050%		0.012%
2	**Trimmed error 5% of Line 1**		**0.003%**		
3	**Temperature drift error (Line 1)**				**0.012%**
	ADC Iteration #2: AD7705, ΔΣ, 16-bit, 4CH, SPI				
1	ADC Offset Error: 0.0 (with internal cal), 0.5µV/°C	0V	0.000%	10µV	0.000%
2	ADC Gain Error: 0.0 (with internal cal), 0.5ppm/°C	0µV	0.000%	25µV	0.001%
3	ADC INL: ±0.003% FSR	75µV	0.003%		
4	**Trimmed error 5% of (Line 1 + Line 2, root sum squares) + Line 3**	**75µV**	**0.019%**		
5	**Temperature drift error (Line 1 + Line 2, root sum of squares)**			**26.9µV**	**0.001%**
	Current Sense Iteration #2				
1	Shunt Error 1Ω: VMP-1R00-1.0-U, 1%, 20ppm/°C	10mΩ	1.000%	400µΩ	0.040%
2	In Amp: AD8223: V_{OS} ±100µv, 1.0µV/°C	100µV	0.005%	20µV	0.004%
3	In Amp Input Voltage Error Total : Line 1 + Line 2	10.1mV	1.010%	420µV	0.042%
4	In Amp Gain Resistor R_G: ERA-6ARB333V(0.1%, 10ppm/°C)		0.100%		0.020%
5	In Amp gain Error (0.02% ±2ppm/°C)		0.020%		0.004%
6	Total In Amp gain error (Line 4 × Line 5)		0.120%		0.024%
7	**Trimmed error 3% of (Line 3 × Line 6)**	**684.8µV**	**0.034%**		
8	**Untrimmable temperature drift error (Line 2 × Line 3)**			**1.3mV**	**0.066%**

Part 9 - overall system design

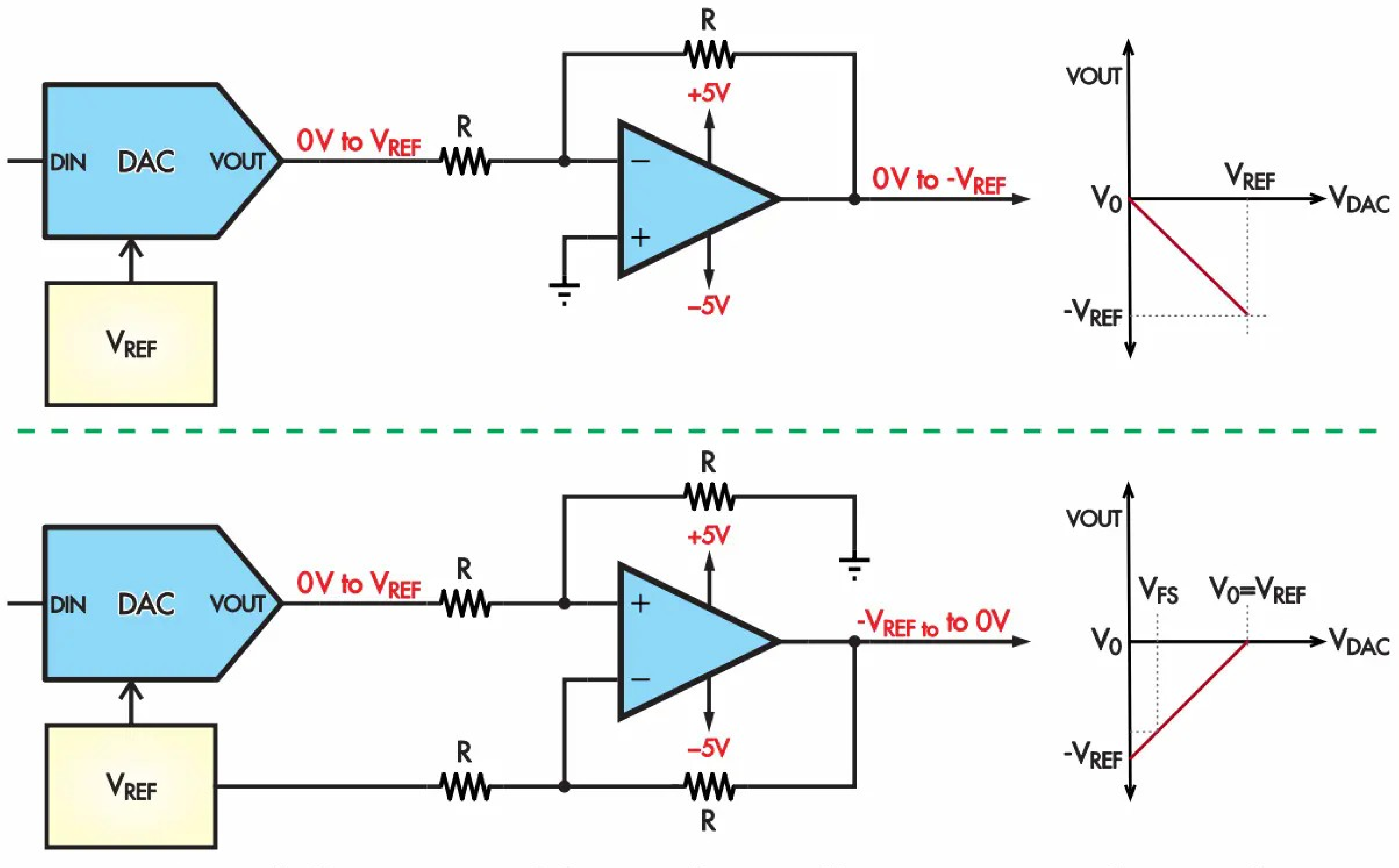


Fig.4: using a difference amplifier as shown allows us to avoid using the very lowest part of the DAC's output range (errors lie there).

Circuit Block	Error @ 25°C	Additional Error ±20°C	Voltage Setpoint	Current Setpoint	Voltage Sensing	Current Sensing
DAC	0.160%	0.010%	1	1		
Setpoint signal conditioning	0.003%	0.030%	1	1		
Summing node	0.010%	0.020%	1	1		
Buffer	0.000%	0.000%	1	1	2	1
Current sensing	0.034%	0.131%		1		1
Voltage divider	0.010%	0.020%			1	
ADC	0.056%	0.080%			1	1
Voltage reference	0.003%	0.012%	1	1	1	1
Target error @ 25°C			0.500%	1.000%	0.100%	0.100%
Target additional error ± 20°C			0.250%	0.500%	0.050%	0.050%
Total @ 25°C			0.175%	0.209%	0.069%	0.093%
Total additional ± 20°C			0.072%	0.203%	**0.112%**	**0.223%**
Total error 5°C to 45°C			0.248%	0.412%	**0.181%**	**0.316%**

Table 3 – we have met the targets for most errors but the voltage and current sensing circuits

You will recall that we need a negative setpoint signal, so we anticipated a signal conditioning block between the DAC and the controller. The simplest way to do this would be to use a unity-gain inverting amplifier, as shown at the top of Fig.4. With a 2.5V reference, this gives us an output in the range 0V to -2.5V, as shown in the graph on the right of the figure.

Unfortunately, this does nothing to help with our near-zero problem.

However, if we use a difference amplifier as shown at the bottom of Fig.4, we shift the DAC output down by V_{REF}, rather than inverting it. The upshot is that we will get very nearly a true 0V output when the DAC is at full scale. As the DAC output approaches zero, the circuit output approaches $-V_{REF}$, corresponding to the maximum output voltage.

By setting the full-scale setpoint voltage to something less than $-V_{REF}$; say, letting -2.0V correspond to a 20V output, we avoid the very bottom part of the DAC's output voltage range and the errors that may lie there.

The downside is that the full-scale DAC code now represents a setpoint of zero and a code near (but above) zero represents full-scale. This inconvenience is easy to remove in the firmware with a simple subtraction.

So, for the sake of the exercise, we will set the voltage setpoint range to 0V to -2.0V representing, 0V to 20V, and the current limit setpoint voltage to 0 to -2.0V, representing 0A to 1.0A.

We can now select a candidate DAC. I chose the MCP48FVB22 low-cost two-channel 12-bit serial DAC from Microchip for this exercise. Its specifications are shown in the appropriate section of the error budget table. Since 2V is our full-scale output, I have used that as the basis for converting between absolute and relative errors.

The upshot is a DAC error of ±0.16% at 25°C with another ±0.01% over the operating temperature range, well inside our ±0.25% budget.

The errors for the setpoint signal conditioning are calculated as we have shown in previous articles. I used a matched resistor array and a low-cost zero-drift op amp, which gives us an overall 25°C error of ±0.003%. In this case, the resistors' ±15ppm temperature drift means we have an order of magnitude more error over the temperature range; however, this is still well inside the error budget.

The summing node errors are dictated by the matching of the resistors – in this case, I used ±0.1% tolerance resistors with ±10ppm/°C drift. The resulting 25°C and temperature drift errors are ±0.01% and ±0.02%, respectively. The buffer uses the same zero-drift op amp as the signal conditioning, and this circuit yields errors that are so low as to be insignificant in our application.

The results for each of these blocks (plus the rest, which I will not discuss in detail) are summarised in Table 3. This is similar in format to the error budget table. The calculated errors for each block are on the left, with the signal chain errors calculated as we did earlier.

All the resulting 25°C signal chain errors are lower than or equal to the targets we set, but the temperature drift errors, and therefore the total errors, for the voltage and current sensing functions (shown in red) are not.

Second iteration

It is not at all unusual to find some problems such as this on the first pass. The process we have gone through – specifically, the detailed error calculation spreadsheet – makes it easy to spot the problem areas.

Table 4 – by addressing critical areas, we have achieved our targets for all signal chains

Circuit Block	Error @ 25°C	Additional Error ±20°C	Voltage Setpoint	Current Setpoint	Voltage Sensing	Current Sensing
DAC	0.160%	0.010%	1	1		
Setpoint signal conditioning	0.003%	0.030%	1	1		
Summing node	0.010%	0.020%	1	1		
Buffer	0.000%	0.000%	1	1	2	1
Current sensing	0.034%	0.066%		1		1
Voltage divider	0.010%	0.020%			1	
ADC	0.019%	0.001%			1	1
Voltage reference	0.003%	0.012%	1	1	1	1
Target error @ 25°C			0.500%	1.000%	0.100%	0.100%
Target additional error ± 20°C			0.250%	0.500%	0.050%	0.050%
Total @ 25°C			0.175%	0.209%	0.031%	0.055%
Total additional ± 20°C			0.072%	0.138%	0.033%	**0.079%**
Total error 5°C to 45°C			0.248%	0.348%	0.064%	0.135%

These are the ADC's ±0.08% temperature-dependent error, which contributes to both signal chains, and the current-sensing circuit's temperature-dependent errors.

The error calculation table shows us that ADC error is entirely due to the temperature dependency of the ADC's gain, so the only real solution is to find a better one. The device I originally chose, the ADS1115, costs around $7 and has a gain drift of ±40ppm/°C, which I thought was appropriate for this design.

The table shows us that we need to get the gain drift down to ±10ppm/°C, or ideally lower, if we are to make a meaningful improvement.

The slightly fancier AD7705 is a good candidate. It features automatic calibration that all but eliminates fixed offset and gain errors, and an impressively low gain drift of ±0.5ppm/°C. This ADC costs about twice as much as the ADS1115, so we have to decide whether the improvement gained from the substitution is worthwhile or not.

This is the type of judgement call you will frequently have to make, but with a detailed analysis such as this, we have the tools to decide where to spend money to get the best results. I decided to go for it – an easy decision, because this is just an exercise!

The next area of temperature drift error is the current sensing circuit. The error tables show that most of the error comes from the instrumentation amplifier's gain drift. Initially, I used an INA821 because we used it in an earlier article. However, we could replace it with the similarly priced AD8223.

This has a significantly worse offset voltage (100μV vs 35μV), but much better gain drift (±2ppm/°C compared to ±35ppm/°C). Replacing the op amp reduces the current sense circuit's temperature error from 0.22% to 0.08%.

This illustrates another type of trade-off we sometimes encounter – trading off one specification against another in similarly priced chips. Building an error budget like I have done here is extremely helpful in working out which parameters really matter for your application.

So, with the changes we have made in iteration 2 (Table 4), we have exceeded our target 25°C specifications by a factor of about two across the board. We have also met the total error over the temperature range across the board, even though the drift figure for the current sense circuit is still a bit higher than the target.

This is another useful lesson – while we have to set targets for both fixed and variable errors, we can trade off underperformance in one with overperformance in the other.

Conclusions

With this, we have reached the end of the Precision Electronics series. If I have one closing message for the prospective precision circuit designer, it would be that a bit of time spent at the beginning with a pile of data sheets and a spreadsheet will be paid back many times over when it comes to building and validation of your designs.

I am as keen as anyone to lay out a board, get my hands on a prototype and sit down at the bench, but I have learned from experience that if I skip the homework, I will pay for it later in frustration and avoidable rework! PE

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A complete kit is available from Silicon Chip – see page 78 for details.

COMPACT HIFI HEADPHONE AMPLIFIER

Introduced last month, our new Stereo Headphone Amplifier fits in a neat package and has two sets of inputs with individual volume controls. Having described its performance and how it works, we'll go over some notes on the PCB design before getting into construction and testing.

The Headphone Amplifier circuit is fairly basic and uses all low-cost and common parts, but it delivers great performance in a small package. It's suitable for relative beginners, with nothing being terribly tricky during the assembly process. Despite that, it still gives a very professional result.

PCBs for hifi circuits are always a bit challenging to design due to the tiny levels of distortion and interference that are required to achieve good performance. So let's take a brief look at what was involved in designing this one.

PCB design

It was a little tricky to fit everything into a relatively small (148 × 80mm) PCB using through-hole components, but we managed that, and the result is shown in Fig.8 and the photos. The power supply section has been kept on the left side, with the input section in the middle and the amplifier section on the right.

The incoming signals arrive at the RCA connectors at the top of the board, flow down through the filtering and coupling components to the buffer op amps at lower middle, then to the volume control pots. They go to the mixer op amp to the right, and up to the transistor buffer section above, then right to the output filter and down to the output sockets.

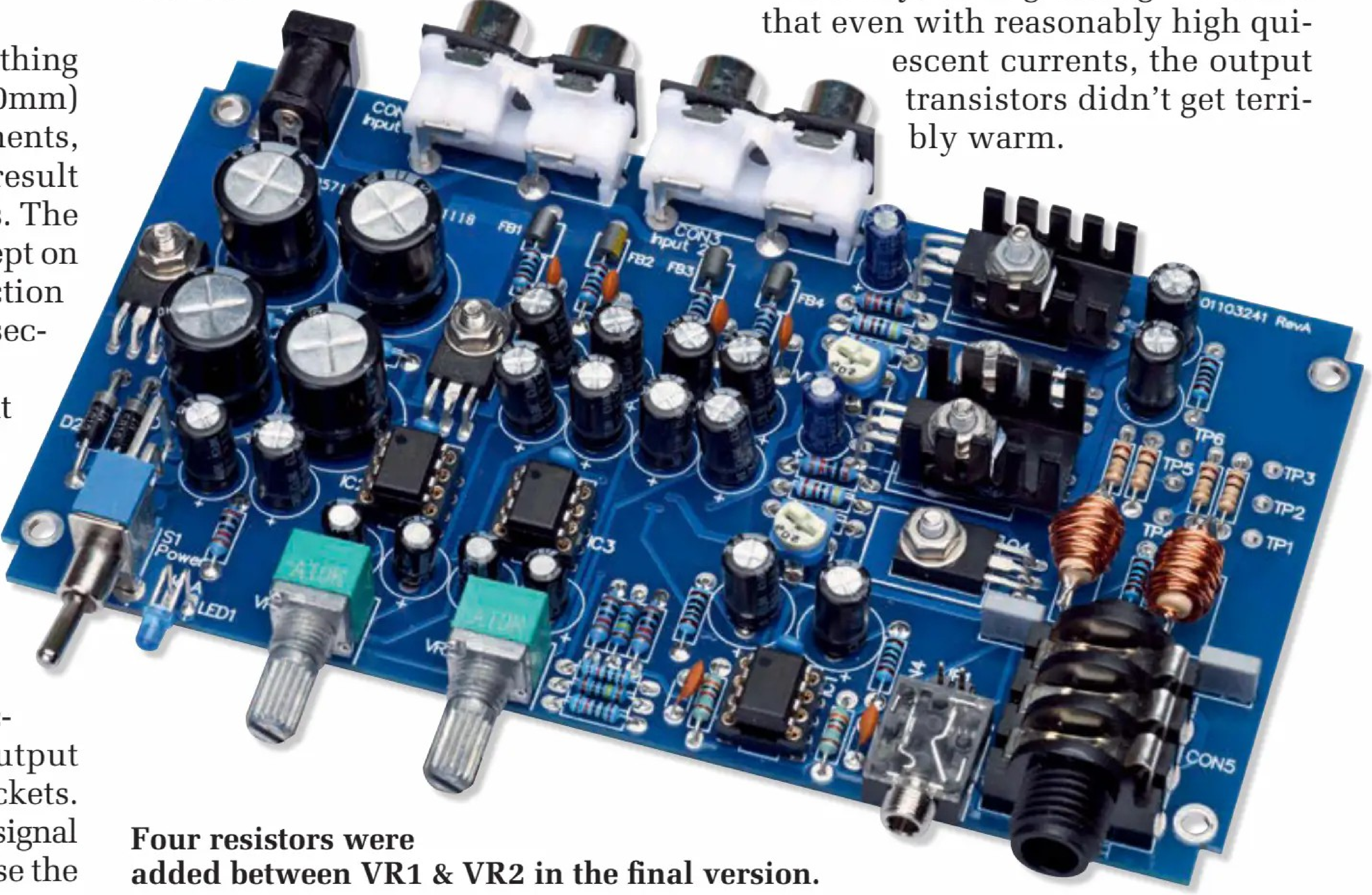
This arrangement keeps all the signal tracks relatively short, to minimise the chance of picking up EMI or magnetic/electric fields from other parts of the PCB. It also keeps the component arrangement neat.

As power needs to flow from the supply on the left side to the transistors at upper right, the positive and negative supply tracks are kept fairly wide and close together so that the magnetic loop is small. That reduces the amount of supply-ripple-induced distortion entering the sensitive signal tracks in the middle of the PCB.

The output transistors have local 100μF bypass capacitors (shared between the channels) to help reduce the effect of the resistance and inductance of those supply tracks. All major ground returns are kept separate back to the power supply common point (similar to star Earthing) so that half-wave rectified currents don't get into the signal grounds and increase distortion.

If you're wondering why only the NPN output transistors have small heatsinks attached, it definitely isn't because we didn't check whether there would be enough room for all four heatsinks to fit side-by-side on the PCB!

Actually, during testing we found that even with reasonably high quiescent currents, the output transistors didn't get terribly warm.



Four resistors were added between VR1 & VR2 in the final version.

Still, as there was room to fit small heatsinks to the NPN output transistors (Q3 & Q5), we did so. That's because these transistors have the V_{BE} multipliers (Q7 & Q8) mounted on top, so they won't be able to dissipate heat as effectively as the PNP output transistors (Q4 & Q6) will.

Also, the PCB is designed to draw heat away from all the transistors that are mounted on it (including those in the power supply). However, as the PCB's ability to absorb, distribute and radiate heat is limited, we figured that by keeping Q3 & Q5 cooler with small heatsinks, that will reduce the total heat load on the board and thus effectively improve cooling for Q4 & Q6 as well.

The heatsinks are actually sandwiched between each NPN output transistor and its associated V_{BE} multiplier transistor, with thermal paste in between. As the thermal resistance of the heatsink is low, that shouldn't have any significant impact on thermal tracking for the V_{BE} multipliers.

While we're on the topic of output transistor ratings, we also need to keep in mind their continuous current limits of 1.5A each, especially during plugging and unplugging headphones. The output transistors have an h_{FE} (current gain) of around 50 times at their limit of 1.5A, regardless of the junction temperature.

That means, to exceed their 1.5A current limit would require a base drive of over 30mA (1.5A ÷ 50). While the NE5532 data sheet says it can typically source or sink 38mA, that's with a ±15V supply and under short-circuit conditions.

In practice, due to supply droop and other factors, with our recommended 9V AC plugpack, we were unable to get our prototype to get anywhere near the limit. Having said that, we didn't deliberately short-circuit the output, so we can't promise it's short-circuit proof. But we think, if you are careful not to abuse it, it should be OK.

Construction

The Headphone Amplifier is built on a double-sided PCB coded 01103241 that measures 148 × 82mm. The same PCB is used regardless of which version you are building. Fig.8 is the component overlay diagram that includes all components for building the full version of the Amplifier, with two sets of stereo inputs.

Fig.9 shows the same arrangement as Fig.8 but without the two buffer op amps. If you're building it from a kit, you might as well build the full version as they are included, but it is possible to leave those two op amps out and save a few dollars. There will be more interaction between the volume controls, though.

Fig.10 shows the PCB with just the components needed for one stereo input. We've chosen to retain CON2, but you could keep CON3 instead and fit the resistors, capacitors and potentiometers in the positions to the right instead.

Regardless of which version you're building, start by fitting all the smaller (¼W and ½W) resistors. They have colour-coded stripes that you can decode with the aid of the table in the parts list. Still it's safer to check each set's value with a DMM set to measure ohms before installing them.

All the smaller resistors are laid flat on the PCB, so bend their leads, insert them, solder them and trim the excess. For the four 100Ω resistors, slip a ferrite bead over one of the leads before inserting it into the board. Solder the shorter end, then pull the other lead with a pair of pliers so it's tight before soldering it. That should stop the ferrite bead from rattling if you move the board.

Next, solder the two diodes, which are the same type. Make sure that both have their cathode stripes facing up, towards Q2.

If using IC sockets, solder them in place now, ensuring the notches all face up as shown on the overlay diagrams. Otherwise, solder the ICs directly to the PCB, again ensuring that their notch or pin 1 dot faces up.

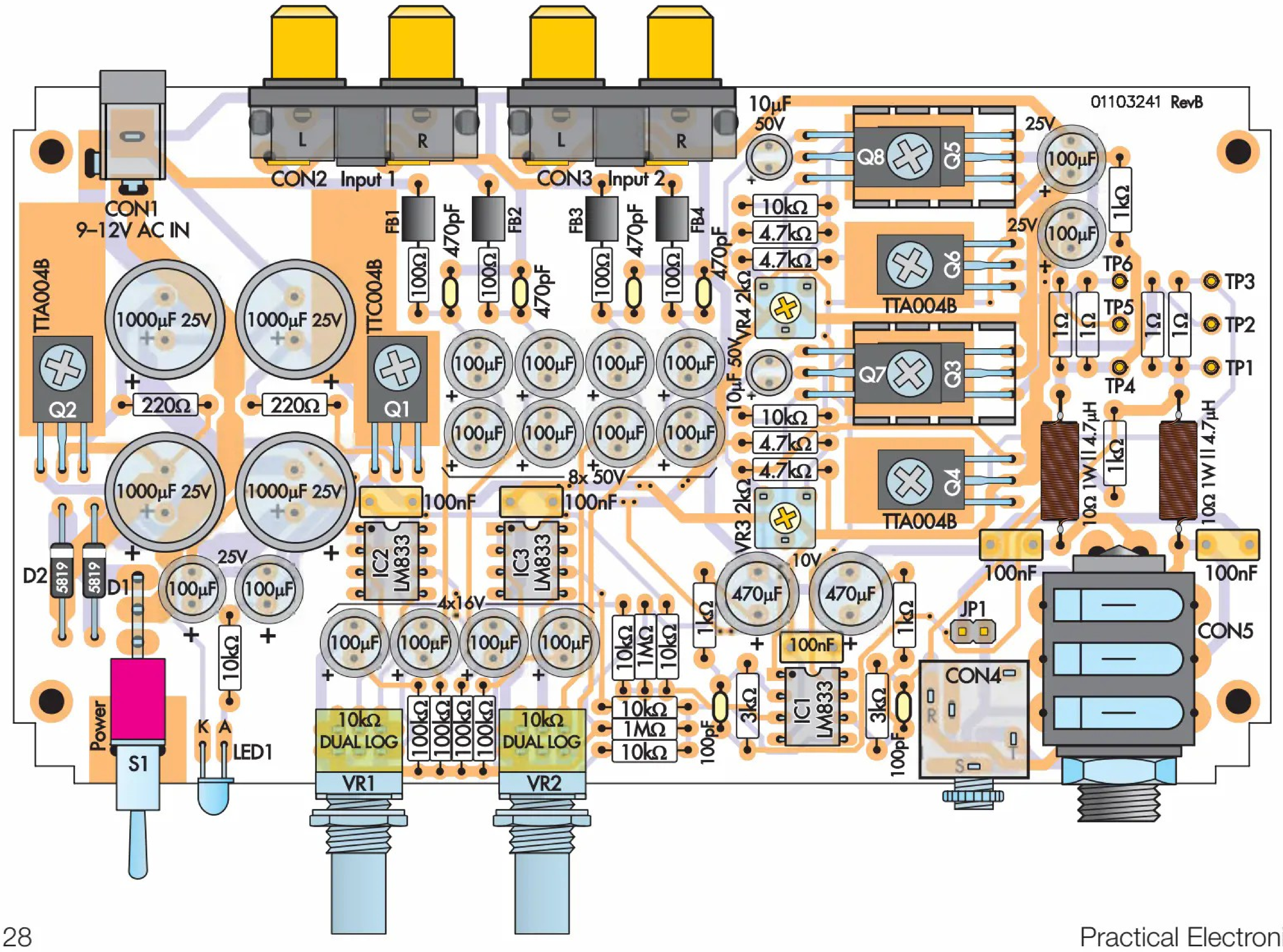
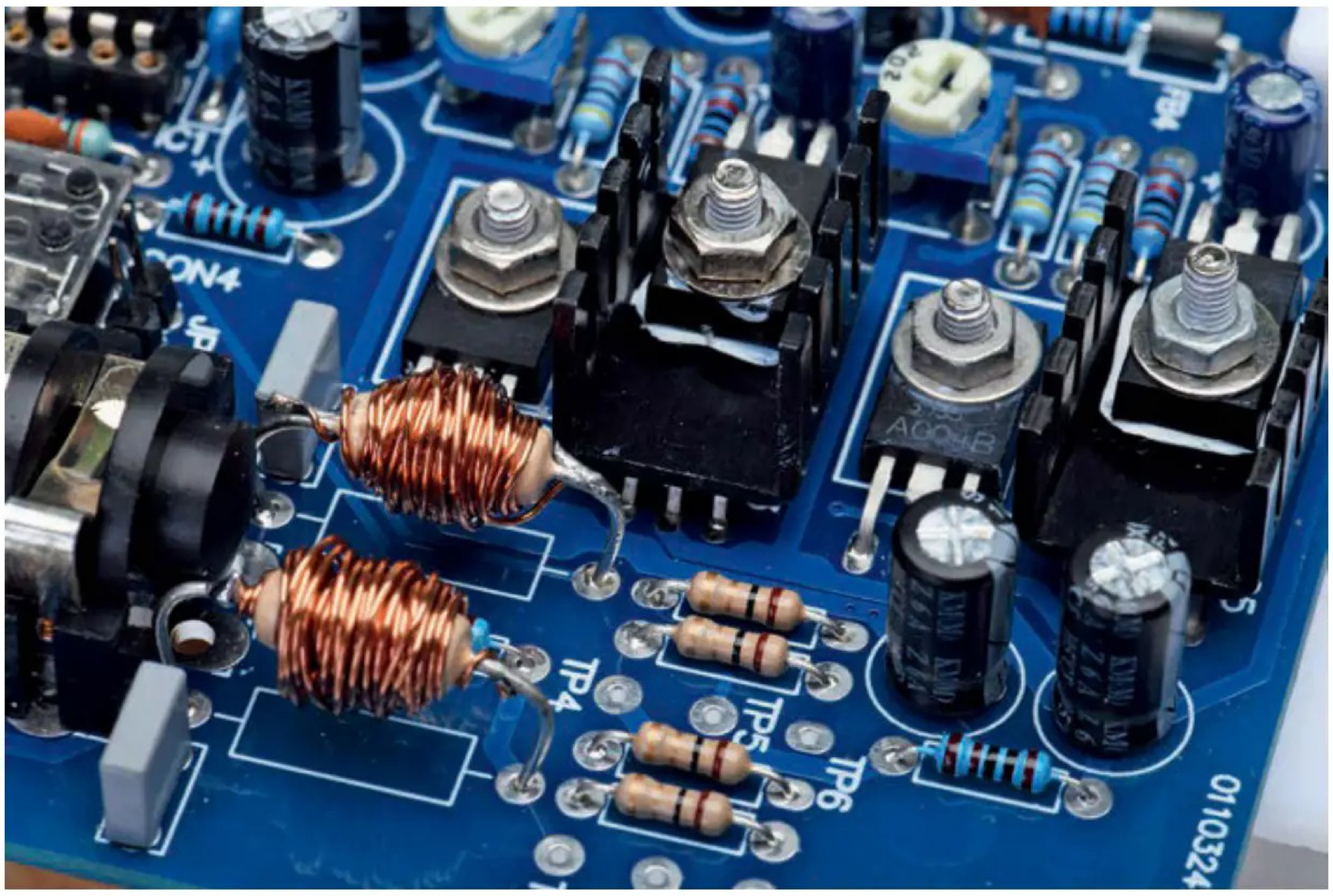


Fig.8: use this overlay diagram as a guide to where to mount each component. This shows the full version with two buffered stereo inputs. Don't forget to add the ferrite beads to the 100Ω resistors before soldering them and watch the orientation of the diodes, ICs and electrolytic capacitors.

Compact Hi-Fi Headphone Amp pt2


CON4
TP4
TP5
TP6
01103241

The output filter inductors are wound on the bodies of the 1W resistors they're paralleled with. You could add heatshrink tubing on top if you want.

This is important as they won't work if reversed!

Now is a good time to fit CON4 if you are using it. Once its pins are lined up with the pads, it should slot right into place. Solder it flat on the PCB.

Next mount transistors Q2, Q4 & Q6. These are all the TTA004B PNP type. Make sure the writing is on the top side, then bend the leads down a few millimetres from their bodies so that they fit through the PCB pads while the mounting hole on the tab lines up with the one on the board.

Add a small amount of thermal paste to the underside of each transistor, then feed a 10mm M3 machine screw up from underneath and push the transistor body over its shaft. Add a flat washer and hex nut on top and tighten while stopping the transistor body from rotating. Check the body is aligned properly, then solder and trim the leads.

Use the same procedure to fit Q1, which is a TTC004B. Leave the other transistors off for now.

Next, mount the two trimpots. They are the same type and only fit one way. Then move on to the capacitors, starting with the ceramics, which are not polarised, so they can go in either way around.

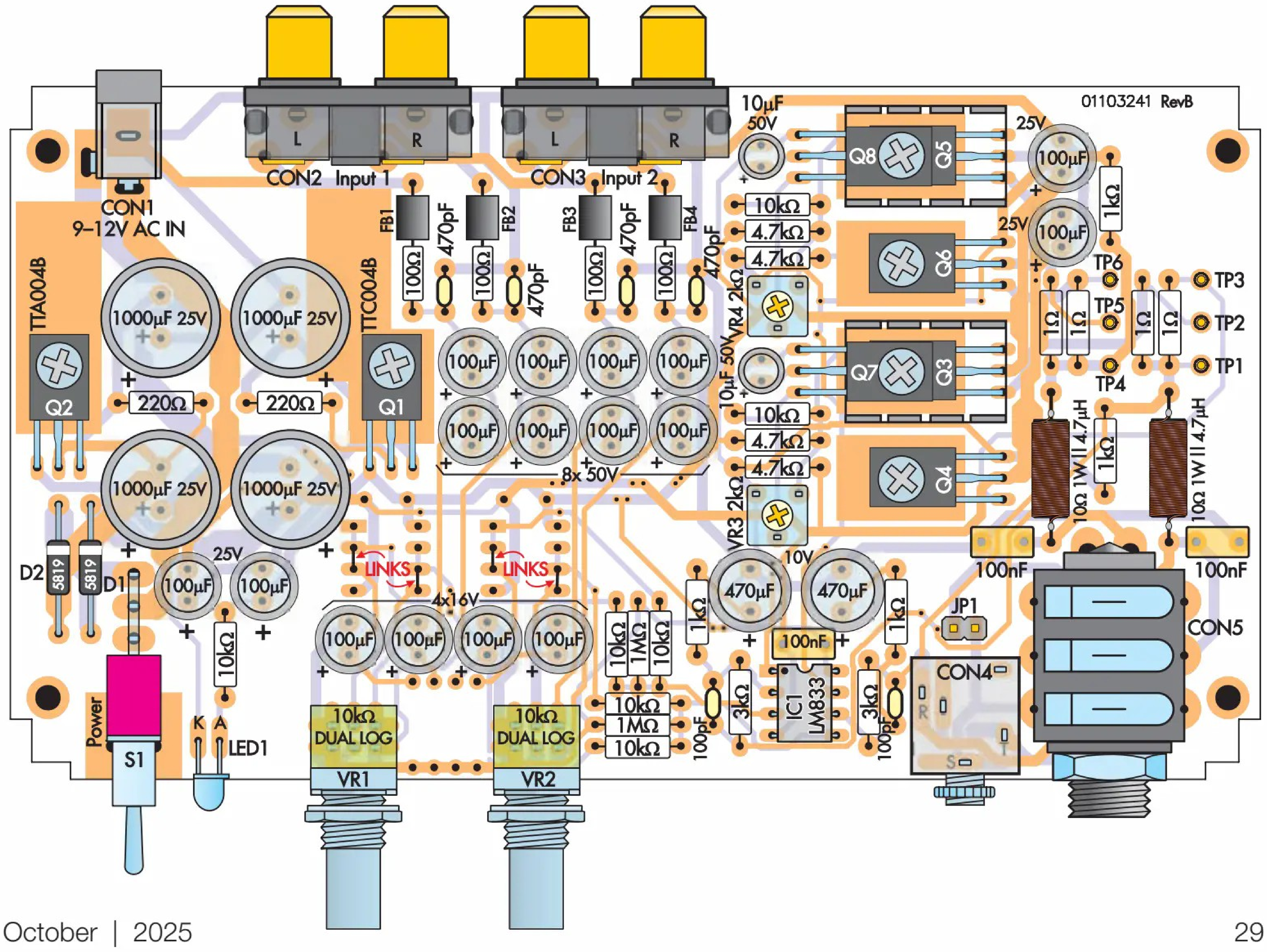
Two of the 100nF capacitors are recommended to be MKT types; fit them next. They are also unpolarised. The other 100nF capacitors can be MKT, ceramic or multi-layer ceramic, none of which are polarised.

Then move on to the electrolytic capacitors, all of which are polarised. In each case, the longer lead goes into the pad next to the + symbol, with the stripe on the can facing the opposite way. The only thing to watch out here, apart from the polarity and the values being correct, is that there are three different types of 100μF capacitors specified.

The four or eight capacitors marked 50V (in the middle of the board) should ideally be 50V types, to make the inputs as robust as possible. They could be lower-rated (eg, 35V) if absolutely necessary. The two low-ESR 100μF capacitors in the power supply section and two more at upper-right must be rated at least 25V, although higher-voltage types are suitable if they will fit.

The two/four 100μF capacitors near VR1/VR2 can be 16V types, although a

Fig.9: here are the difference if you're building the two-input version without the buffer op amps. Fit the four links instead of the ICs and leave off the four 100kΩ resistors.


CON1
9–12V AC IN
CON2 Input 1
CON3 Input 2
L
R
L
R
10μF 50V
01103241 RevB
TTA004B
Q2
TTC004B
Q1
1000μF 25V
1000μF 25V
1000μF 25V
1000μF 25V
220Ω
220Ω
FB1
FB2
FB3
FB4
470pF
100Ω
100μF
8x 50V
10kΩ
4.7kΩ
4.7kΩ
VR4 2kΩ
VR3 2kΩ
10μF 50V
Q8
Q5
Q6
Q7
Q3
Q4
25V
100μF
1kΩ
TP1
TP2
TP3
TP4
TP5
TP6
1Ω
10Ω 1W II 4.7μH
100nF
CON5
D1
D2
5819
25V
10kΩ
LINKS
4x16V
10V
470μF
1MΩ
3kΩ
100pF
IC1 LM833
JP1
CON4
Power
S1
K A
LED1
10kΩ DUAL LOG
VR1
VR2

higher rating certainly won't hurt, as long as they will fit.

Now is a good time to solder the two-pin header for JP1 in place. After that, fit VR1 and/or VR2, making sure they are pushed fully down and their shafts are perpendicular to the edge of the PCB. Also fit the barrel socket, again making sure it is straight and flat before using generous amounts of solder to attach it.

The RCA sockets need the projection on the top cut off. It's easiest to do it before mounting them on the board. Use a hacksaw or rotary tool to cut them off in line with the top edge of the socket face, then file or sand off any burrs or projections. Snap them into the PCB and make sure they're flat before soldering the pins.

Similarly, mount the on/off switch next. The LED goes next to the switch, with its lens at the same height as the switch shaft. Bend its leads by 90° about 3mm from the band of the lens, ensuring that when it's inserted into the PCB, its longer (anode) lead will be to the right, as shown in the overlay diagrams. Insert and solder it so that its lens is at the same height as the switch and pot shafts.

Heatsinks

All four remaining transistors are TTC004Bs, and they are held to the board using 15mm or 16mm long M3 machine screws. This bit can get a little fiddly and messy, so keep a damp cloth on hand, along with needle-nose pliers and angled tweezers. The mounting arrangement is depicted in Fig.11.

First, bend the leads of all four transistors down so that they will fit into the PCB pads with the tab mounting hole in the correct position and the writing on the top. Make sure they can be inserted easily and that the tab hole is properly aligned, as that will make the rest of the job much easier.

Insert a machine screw up through the bottom side of the PCB, then add a thin layer of thermal paste on **both sides** of one of the transistors. This will be Q3. Insert its leads and push it most of the way down to the PCB, then add a heatsink over the top, with the longer section projecting to the right (over Q3's leads).

Next, add thermal paste to the bottom side (only) of another transistor and add it on top of the heatsink (Q7). Place a flat washer over the screw shaft, then do up a nut on top. Hold the transistor bodies steady as you tighten the nut, then solder and trim all six leads. Repeat for the other transistor pair.

Winding the inductors

We used 0.4mm diameter enamelled copper wire (ECW) to wind the inductors, although you could use a smaller diameter (down to about 0.25mm) if you happen to already have it. Cut it into two 1m lengths, then use a sharp hobby knife or emery paper to strip the insulation off the ends by 2-3mm.

The inductors are wound using the bodies of the 10Ω 1W resistors as formers. Clamp a resistor in some sort of holder (we used the type that has mini grabbers), then add some solder to the leads on either end of the body. Solder one end of the ECW to that point, with the rest going past the body, then start winding it around the body.

Try to keep it neat and closely spaced at first, although it's basically impossible to keep it neat after the first layer. The good news is that there aren't a huge number of turns required, so it hopefully won't end up a jumbled mess by the time you have finished. Keep it wound tightly around the body, then solder the remaining stub close to the other end of the resistor body.

Use a DMM to measure the resistance across the resistor. It should have dropped to around 0.2Ω (depending on your DMM lead resistance). If it's close to 10Ω, that suggests the solder joint at one end (or both) is bad, so fix it.

Repeat for the other resistor, then bend the leads, insert them into the PCB and solder them at similar heights. There's no significant dissipation in

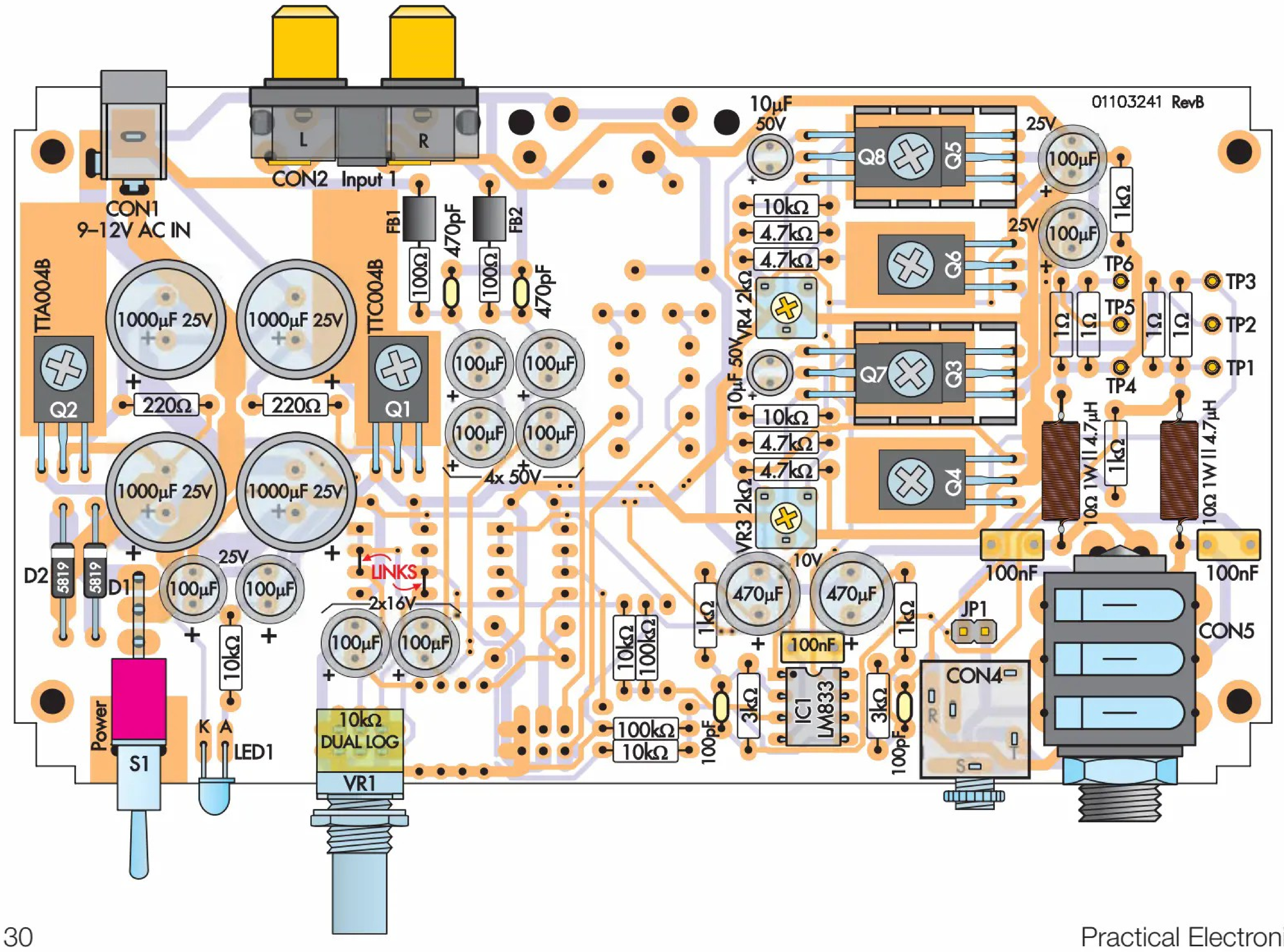


Fig.10: if building the single-channel version, you can leave off either channel; here we're showing CON2 fitted and CON3 not. Only one IC needs to be linked out in this case. In place of the two 1MΩ resistors, use 100kΩ instead.

these devices, but it's easier to solder them spaced off the PCB, so you might as well.

Finally, if you're using the 6.35mm jack socket, CON5, solder it now. It will need to be a low-profile version to fit in the case. Some variants space the body of the socket off the PCB, so make sure that the top of the socket body is no more than 11mm above the board. Push it down fully and solder it in place using generous amounts of solder for good mechanical retention.

Testing

Adjust VR3 & VR4 to be fully anticlockwise and ensure switch S1 is in the up (off) position. Plug in the plugpack and switch it on at the mains. Nothing should happen since the switch is off.

Set your DMM to alternating current (AC) measurement mode (not DC!) in the amps range and connect the probes appropriately. Hold one against switch S1's pad that's closest to the large capacitor (ie, the one at the back & top). While watching the multimeter, touch the other to the middle pad for S1 for a second or two.

If you've used IC sockets and the chips are not inserted, you should see a current draw of only a few tens of milliamps at most, and LED1 should light up. If all three op amps are soldered to the board, the current draw will be closer to 150mA. If you have fewer op amps installed, it will be in between (~50mA for one and ~100mA for two).

If the current draw is a lot higher than that, or LED1 doesn't light up, you have a problem. Disconnect the power supply and check the board for faults like pads bridged with solder, incorrectly orientated components, components in the wrong location etc.

If it seems OK, set your DMM to measure DC volts and hold the black probe to a convenient ground point, such as the left-most pin of JP1 or the bottom-most end (closest to the PCB edge) of one of the row of four 100kΩ resistors between VR1 & VR2. Hold the red probe on pin 8 of one of the ICs and switch the power back on.

You should get a steady DC voltage reading of around 13V DC for a 9V plugpack or 17V for a 12V plugpack. Then touch the red probe to pin 4 of the same IC, and you should get a negative voltage of a similar magnitude.

Next, check the AC voltages at those two pins. The reading should be no more than about 10mV AC (our prototype measured almost exactly 10mV with the ICs in-circuit).

If you are using IC sockets and haven't inserted them yet, switch off the power and wait for LED1 to extinguish. Install all the ICs you require, ensuring that pin 1 goes towards the upper-left corner, near the notches on the sockets.

Now measure the DC output voltages relative to ground. They are available at the bottom ends of the two 10Ω 1W resistors that have the ECW wrapped around them. Measure those points relative to ground with the power on (see earlier for convenient ground points) and confirm that the readings are under 50mV (with either polarity). Our prototype measured around -25mV on both channels.

If they are much higher than that, something is wrong, so switch off and search for faults. Rectify any problems you find and re-check the output voltages to verify they are under ±50mV before proceeding.

Adjustment

Connect a DMM set to measure millivolts between TP1 and TP2. The reading should be close to zero initially. Slowly rotate VR3 clockwise and by the time it reaches its midpoint, the voltage reading should start to rise. Adjust it for a reading close to 25mV (meaning 25mA quiescent current). Move the probes to TP1 & TP3 and the reading should be similar.

Now connect the probes between TP4 & TP5 and perform the same adjustment using VR4. You can then check that the reading is similar between TP4 & TP6.

At this point, you are ready for a listening testing. Switch off the power, rotate VR1 and VR2 fully anti-clockwise and plug headphones or earphones into one of the sockets. Don't put them over or onto your ears yet. Connect a low-level stereo audio signal source to one of the inputs, cue it up and switch the amplifier back on.

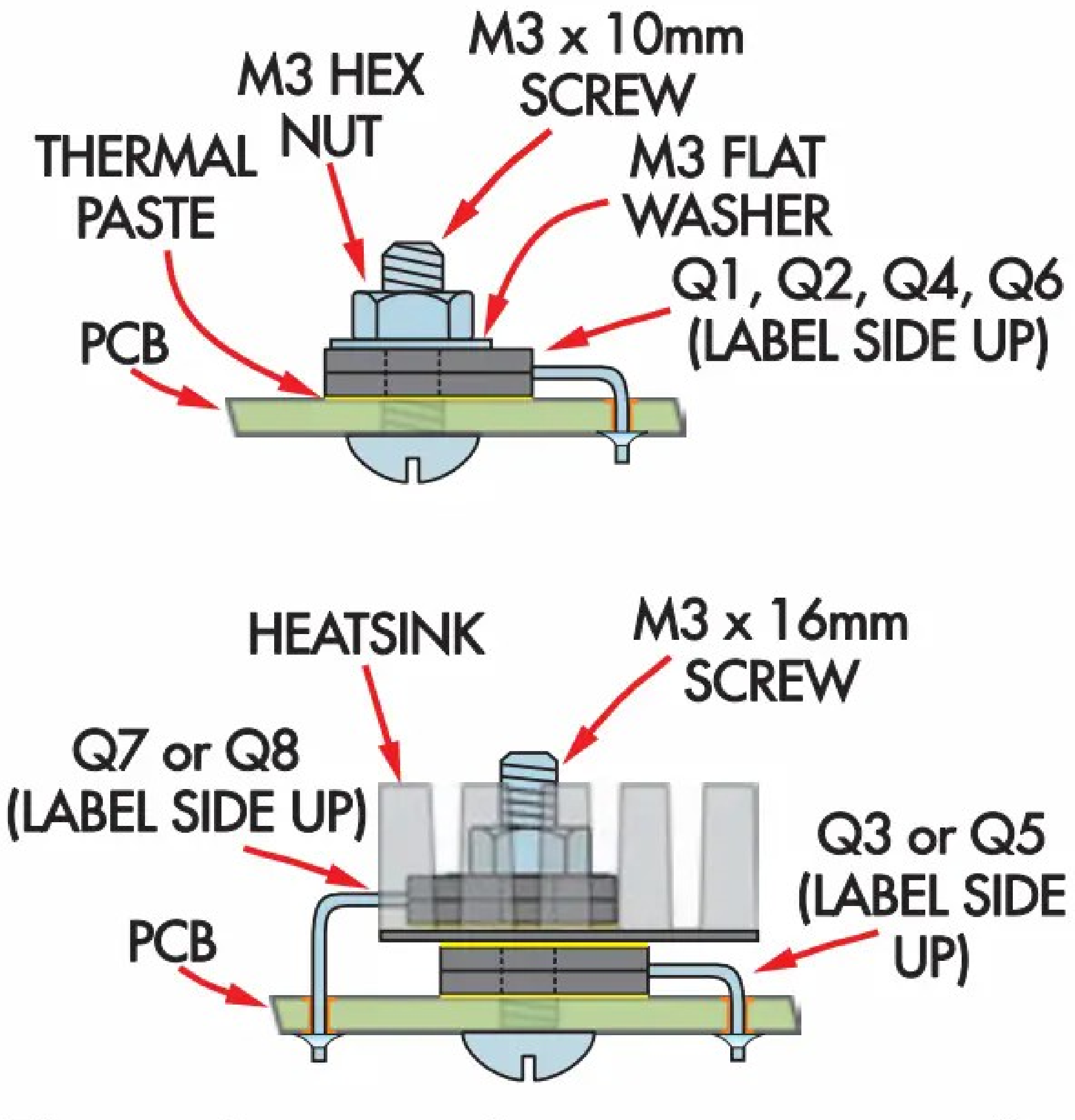


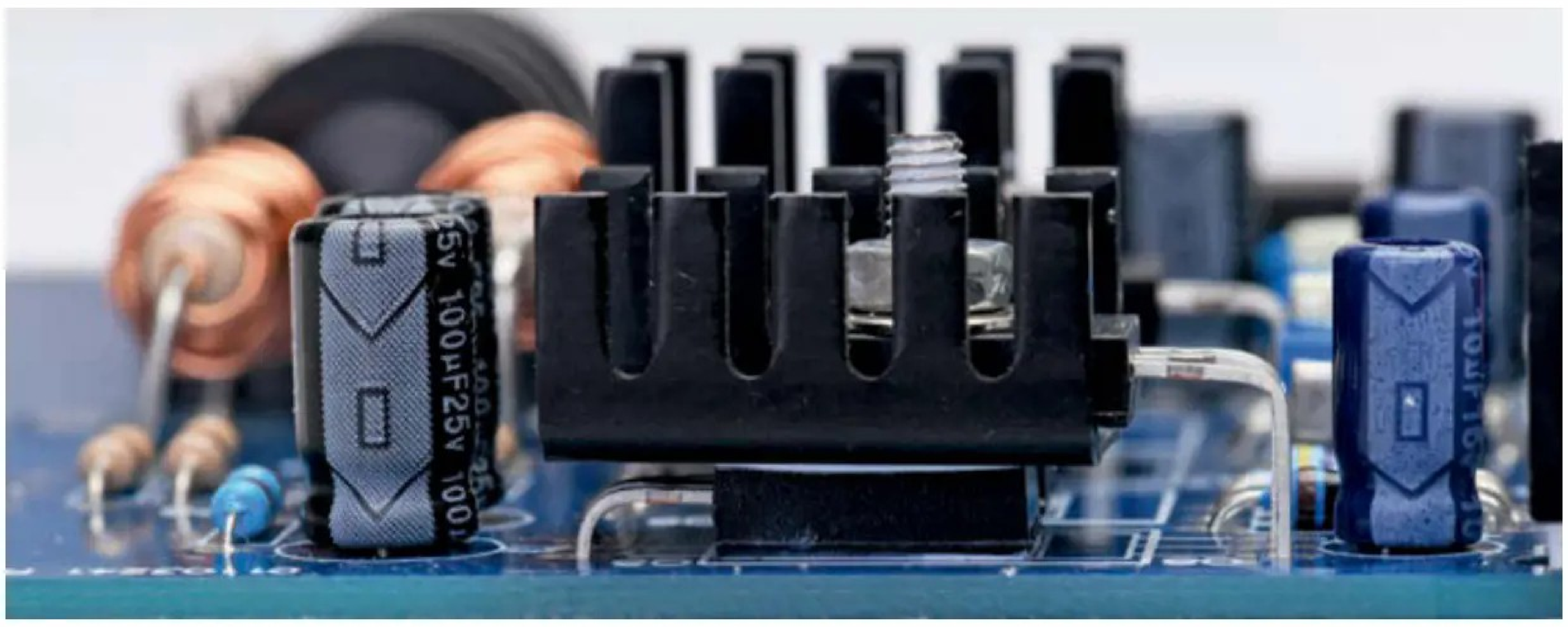
Fig.11: the mounting arrangements for the power transistors and heatsinks.

Slowly wind up the volume pot associated with the channel you're using (VR1 for CON2 or VR2 for CON3) and check that you can hear audio by moving the headphones/earphones closer to your ears. If it sounds normal, try putting them over/into your ears and adjust the volume to a comfortable level. Verify that the audio sounds normal and undistorted, with similar levels for both channels.

If it sounds strange, switch off and look for faults on the PCB.

Jumper option

Before assembling the case, decide if you want to put a jumper shunt on JP1. With it out, if you plug headphones into both sockets, audio will only come from CON5 (CON4 will be disconnected). With it in place, the headphones will be connected in parallel and both will get audio (but possibly not at the same volume!).



A close-up photo of the way the heatsinks are fitted. This is from the opposite side to that shown in Fig.11.

If you've only fitted CON5, it doesn't matter if you put a jumper on JP1. If you've only fitted CON4, you must add the jumper, or it won't work.

While CON4's ground is disconnected without JP1 if a plug is inserted in CON5, due to the way headphones are wired, you might still get some sound out of headphones still plugged into CON4. It's unlikely to be anywhere near full volume, though. If it bothers you, simply unplug the unused pair.

Case preparation & installation

Preparing the case is relatively straightforward: all the holes to be made are in the front and rear panels, and they are all round, so you can use a drill (a stepped drill bit makes it easier). The locations are shown in Fig.12. There are six holes to make in the front panel and five at the rear, from 3mm to 10mm in diameter.

You can download a PDF of Fig.12 from **siliconchip.au/Shop/11/7406**, print it out at actual size, cut it out and stick it to the panels using weak glue or scotch tape. Drill small pilot holes as accurately as you can in the centre of each location, then remove the templates and drill them out to the sizes shown.

Deburr the holes and check that the panels fit over the assembled PCB in the case. You may need to slightly enlarge some holes if their locations are not perfect.

The bottom of the case can be identified as it has four small circular recesses for feet. Stick small rubber feet in or near those locations, then secure the PCB to the base using four small self-tapping screws. Remove the nuts from the jack sockets, slot the lid on top, then push the front and rear panels in place. After that, you can attach the knobs.

Our initial prototype was designed with the potentiometer and socket shafts essentially being flush with the front panel, so we couldn't reattach their nuts. We didn't think that was a problem as it seemed robust enough without them.

Still, we made some adjustments to the final PCB so that the on/off switch, volume control pots and 3.5mm jack socket are closer to the front. That means you should be able to get the nuts back on the pots, which will provide a bit of extra rigidity, and it will make plugging into the 3.5mm socket easier, although you probably won't be able to get its nut on.

We have kept the front of the 6.35mm socket close to being flush with the front panel as we think it's neater, and it's mechanically secure enough without it.

Using it

It's generally a good idea to wind VR1 & VR2 back to zero (or close to it) before playing audio if you don't know if the levels set previously are appropriate. Then slowly advance the volume control associated with the input (VR1 for CON2 & VR2 for CON3) until you reach a comfortable volume level.

It's best to avoid 'live plugging' headphones as they can short the outputs when doing so. It will probably be OK, but it's safer to switch the device off before plugging or unplugging.

We also suggest you remove the headphones/earphones when switching the amp on or off to avoid any painful clicks or pops that may occur. This will also protect you in case you switch it on and the volume level is set too high.

The amp draws no power when switched off, although AC plugpack will draw some power from the mains even when it has no load. So if you want to minimise power consumption when the amp is off, switch off the plugpack at the wall or unplug it when not in use.

If you ever have to get the case apart again, it's a bit tricky but it can be done. Remove the knobs and nuts, then detach the front panel on the switch side. The rear panel is almost impossible to remove once assembled as the RCA sockets prevent you from flexing it in such a way to release the tabs, so don't try.

Once you have the front panel off on one side, pull at the bottom on the jack socket side and squeeze the main part of the case in, and it should pop off. You can then gently lever the top off and pull it forwards to release the rear panel. **PE**

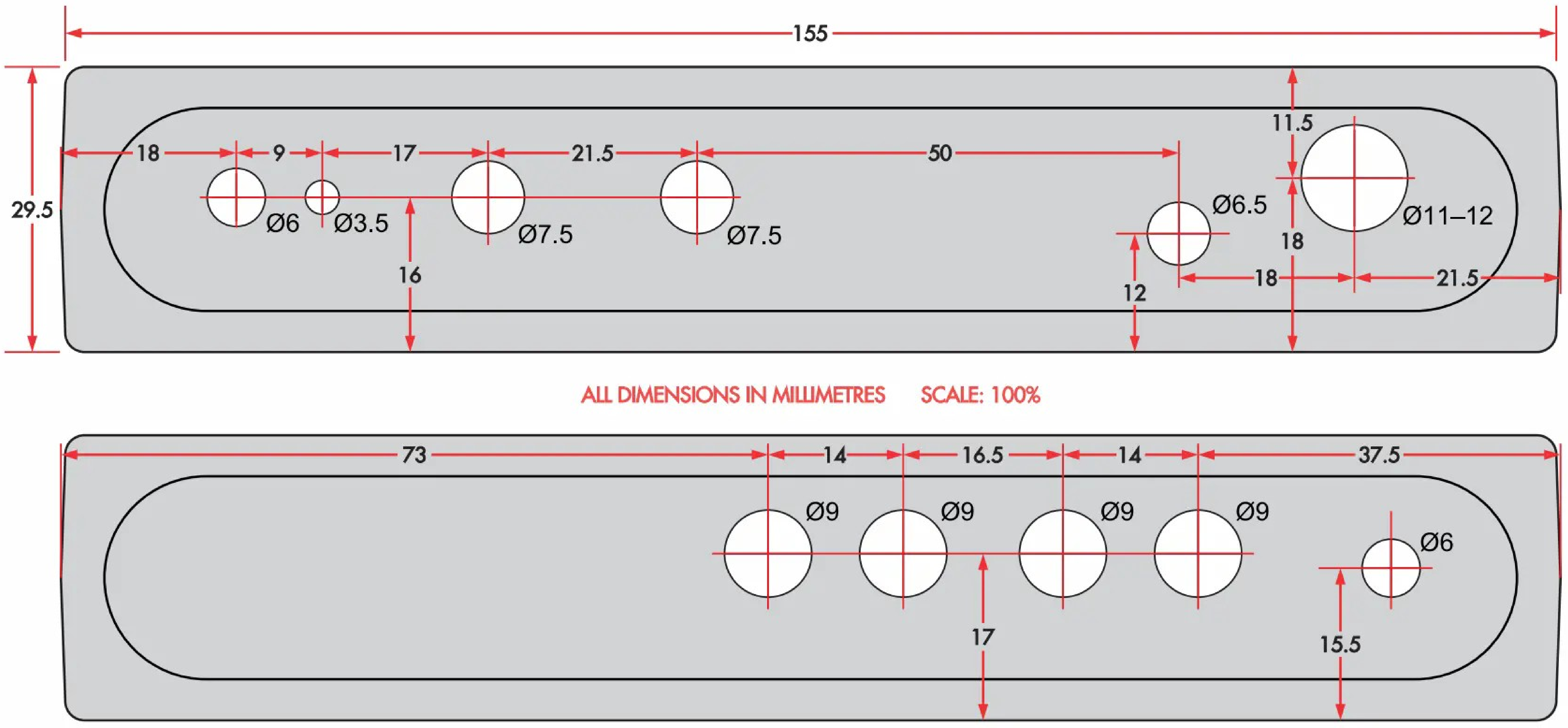


Fig.12: the front (top) and rear (bottom) panel drilling details. Depending on how accurately you drill the holes, you may need to enlarge some slightly before the panels will snap into place. It's best to start them all small and then increase them by a couple of millimetres at a time until they're at full size. If building a single-channel version, only drill the two 9mm holes corresponding to the RCA sockets you have fitted (and the same for the 7.5mm potentiometer holes at the front).

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MAX'S COOL BEANS

By Max the Magnificent

WEIRD & WONDERFUL ARDUINO PROJECTS

Part 10: How good are your reactions?

A **"stream of consciousness" is** when someone writes in a way that directly mimics the way they think, tangents and all. It's basically an unfiltered 'brain dump' onto paper. What's that, you say? You'd like to see an example? Well, this is your lucky day…

As I may have mentioned on occasion, six is my lucky number. I know this because my mum told me so when I was six years old. Talking about air conditioners (we weren't, but we are now), as I pen these words, the heat index outside is 106°F (~41°C). This shows how lucky I am, because it could have been 107°F.

We didn't have air conditioners when I grew up in England. It took us thousands of years to learn how to heat our houses; the thought of trying to cool them down never occured to us. But I now live in America. Fortunately, we have air conditioners. Unfortunately, when I say "we", I fear that this is not in the inclusive sense, because our air conditioner is no longer with us.

Much like the Norwegian Blue in Monty Python's Dead Parrot sketch, we are now the proud possessors of an ex-air-conditioner. Thus, the "outside" qualifier I used earlier was superfluous because conditions are much the same inside. The humidity is 85% (you see, lucky again, it could have been 86%). I'm hot. I'm sticky. And I'm sweaty (and not in a good way).

Also, I've fallen behind, and I must write this column before our illustrious editor turns his fearsome, all-seeing eye in my direction. But it's not all about me (it should be, but it's not); how's your day going?

I must have blinked

A couple of columns ago, we noted that our latest-and-greatest software switch debounce code introduced a delay between our activating a switch and any response (like clocking or clearing our shift register). For reasons we discussed then, we are operating under the principle that our switches may bounce for anything up to six milliseconds (6ms) following their **initial transitions**.

The debounce technique we've adopted waits 16ms following the **final bounce** (to ensure that this **was** indeed the final bounce) before initiating any actions. This means that we may experience a worst-case delay of 6ms + 16ms = 22ms between activating one of our switches and seeing the results.

Having any sort of delay bothers me. However, as I mentioned at the time, this delay is so slight as to be imperceptible. I also stated that, as a 'rule of thumb', we should aim to keep any switch-to-action delays under 50ms.

Of course, this is in the context of human-activated switch events. Things are different when we are talking about switches controlled by (and controlling) machines, like robotic arms on an automated assembly lines, but that's a whole other kettle of parrots.

The reason I used 50ms is that this is what I was told by grizzly old engineers when I was young, naïve, and believed everything I was told. Now I'm older and… well, let's just stick with older. This explains why, in our last column, we decided to test things for ourselves.

We left most of our existing setup as it was. Our only modification to the hardware was to add one LED (and its associated current-limiting resistor), connected to pin 13 of the Arduino (see Fig.1). Even this was optional, as the orange LED on the Arduino is also connected to pin 13.

For the software, we took bits and pieces from earlier programs and munged them together to create our

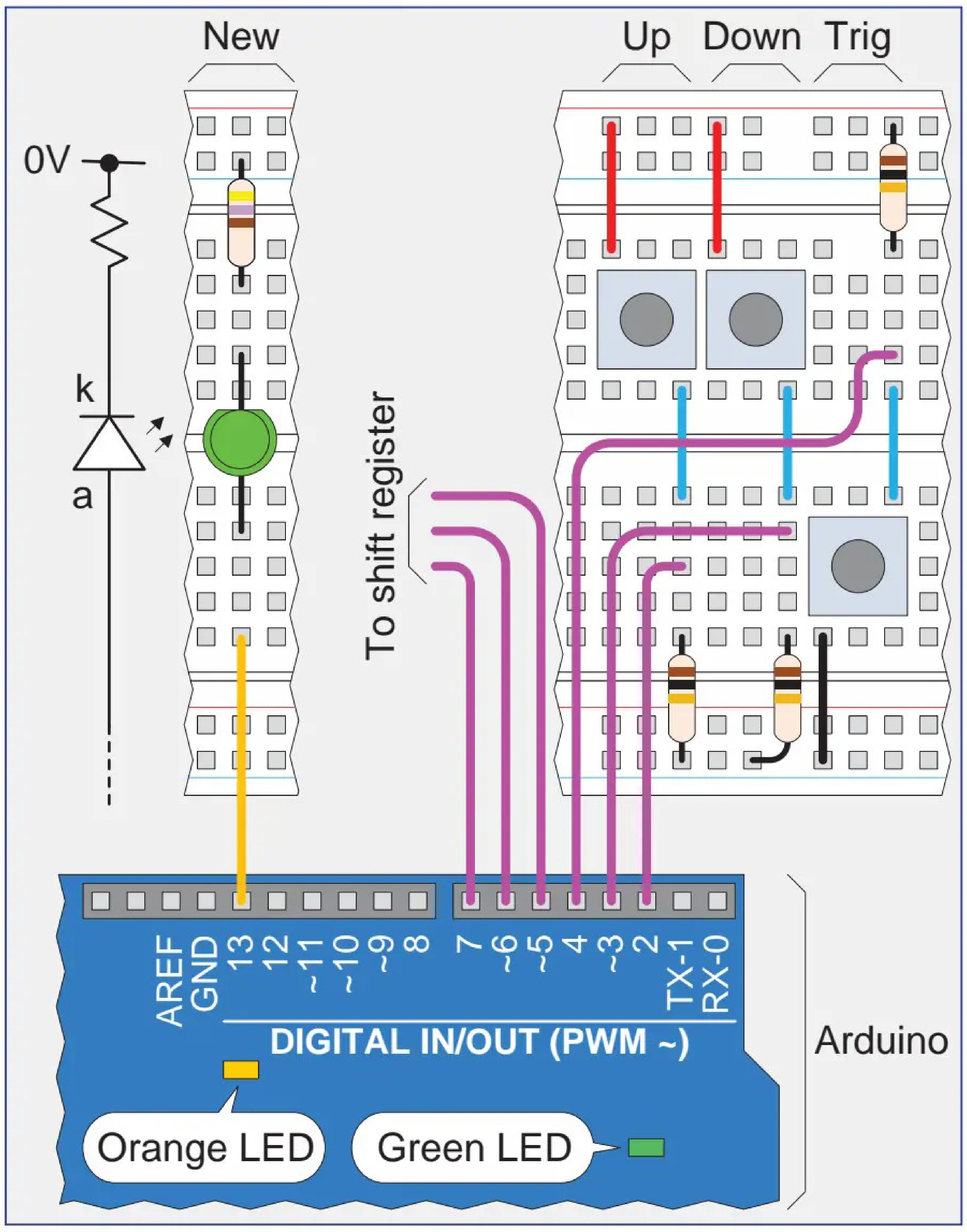


Fig.1: the breadboard configuration of our perception tester.

new perception tester program. You can refresh your memory by downloading this program in the file named **CB-oct25-code-01.txt**. As usual, all files mentioned in this column are available as part of the October 2025 download package from the *PE* website at **https://pemag.au/link/ac7y**

We still use signals from the Arduino (pins 5, 6, and 7) to control the shift register, but only to clear it because we don't want to leave its inputs floating around and have it displaying annoying random values.

The most significant change is in how we handle the signals from the switches to the Arduino (pins 2, 3, and 4). In our program, we set the initial lag (delay) between the switch of interest being pressed or released and the LED changing to 50ms.

We use our left-hand switch to perform the role of 'up', ie, to increase the lag in 5ms steps. We use the middle switch to perform the role of 'down', to decrease the lag in 5ms steps. Whenever we make any such change, we display the current lag value on our host computer's screen using the Arduino's Serial Monitor function.

Last but not least, we use our right-hand button as a trigger to initiate actions. Pressing or releasing this button causes the LED to switch on or off (after the current lag), respectively. Once again, all this was discussed in excruciating detail in our last column.

The interesting part was that before performing this experiment, I honestly believed I'd be able to perceive a 50ms lag, but such was not the case. I was surprised to discover that I can barely discern delays between 150 and 200ms! Is this because I'm ageing like a fine cheese? Did I blink? Or is there more to this than meets the eye (no pun intended)?

More than meets the eye

I was puzzled (which is my natural state), so I decided to conduct some research, by which I mean I asked ChatGPT (I do my best to cross-check everything it tells me, just in case it's having a virtual 'bad hair day'). It turns out that there's a meaningful difference between a user performing this sort of action occasionally versus repeatedly.

In the case of occasional use (single press), which is essentially what we're doing, albeit more than once, the user is more likely to overlook short delays. Since the user isn't in a rhythm, their brain doesn't establish a strong baseline for timing comparison. As a result, delays under about 200ms may seem instantaneous or just 'natural lag', especially if there's no reference point.

By comparison, in the case of repetitive action (rhythmic pressing), our temporal sensitivity increases. The brain adapts and anticipates the response pattern, so any inconsistency or delay becomes more noticeable. Even small lags (like 50ms) start to feel 'off', because the user begins to internally track the expected timing.

As a result, with a 50ms lag, the LED may feel slightly sluggish, while at 200ms, it may feel significantly delayed, possibly frustratingly so when it comes to rhythmic tasks like tapping to music.

Family	Logic Technology / Type	Introduced
74	Standard TTL	1964
74H	High-speed TTL	~1966-1967
74L	Low-power TTL	~1968
74S	Schottky TTL	~1971-1972
74LS	Low-power Schottky TTL	~1975-1976
74AS	Advanced Schottky TTL	~1982
74ALS	Advanced Low-power Schottky TTL	~1983
74F	Fast TTL	~1985
74C	Standard CMOS	~1978
74HC	High-speed CMOS	~1982-1983
74HCT	High-speed CMOS, TTL-compatible	~1983-1984
74AC	Advanced CMOS	~1985-1986
74ACT	Advanced CMOS, TTL-compatible	~1985-1986

Fig.2: the main 74xx-series logic IC families.

Mea culpa

Before we proceed, I'm afraid I may have misled you a little. In these *Weird and Wonderful* columns, we're reusing many of the components that we acquired as part of our *Arduino Bootcamp* series.

As you may recall, we've been using the 74xx series of integrated circuits (ICs) for our experiments. This is an extensive family with numerous subfamilies, each offering different characteristics, such as speed and power consumption.

We began by purchasing a kit of popular 74LS chips (**https://pemag.au/link/ac2k**). We opted for LS (low-power Schottky) parts because this was the dominant subfamily during the late 1970s and 1980s.

Since they were produced in bulk and are now somewhat outdated for modern commercial designs, LS parts are often available at a very low cost, making them attractive for hobbyists, educators, and retro-computing fans.

Later, we added some 74xx595 8-bit shift registers. Since it's harder to track down LS versions of these little rascals, we opted for HC (high-speed CMOS) parts (**https://pemag.au/link/ac1n**). This is where the (potential) problem lies.

I received an email from a reader we shall call William (because that's his name), who asked, "Are there any issues using the 74LS logic gate outputs to drive the 74HC shift register inputs?"

Oh, William, why doth thou vex me so? Unfortunately, William is correct to raise his querulous concern, because there are indeed potential problems.

The approximate chronological order of introduction for the main 74xx00 subfamilies (both TTL and CMOS) is illustrated in Fig.2. TTL stands for transistor-transistor logic, which utilises bipolar junction transistors (BJTs).

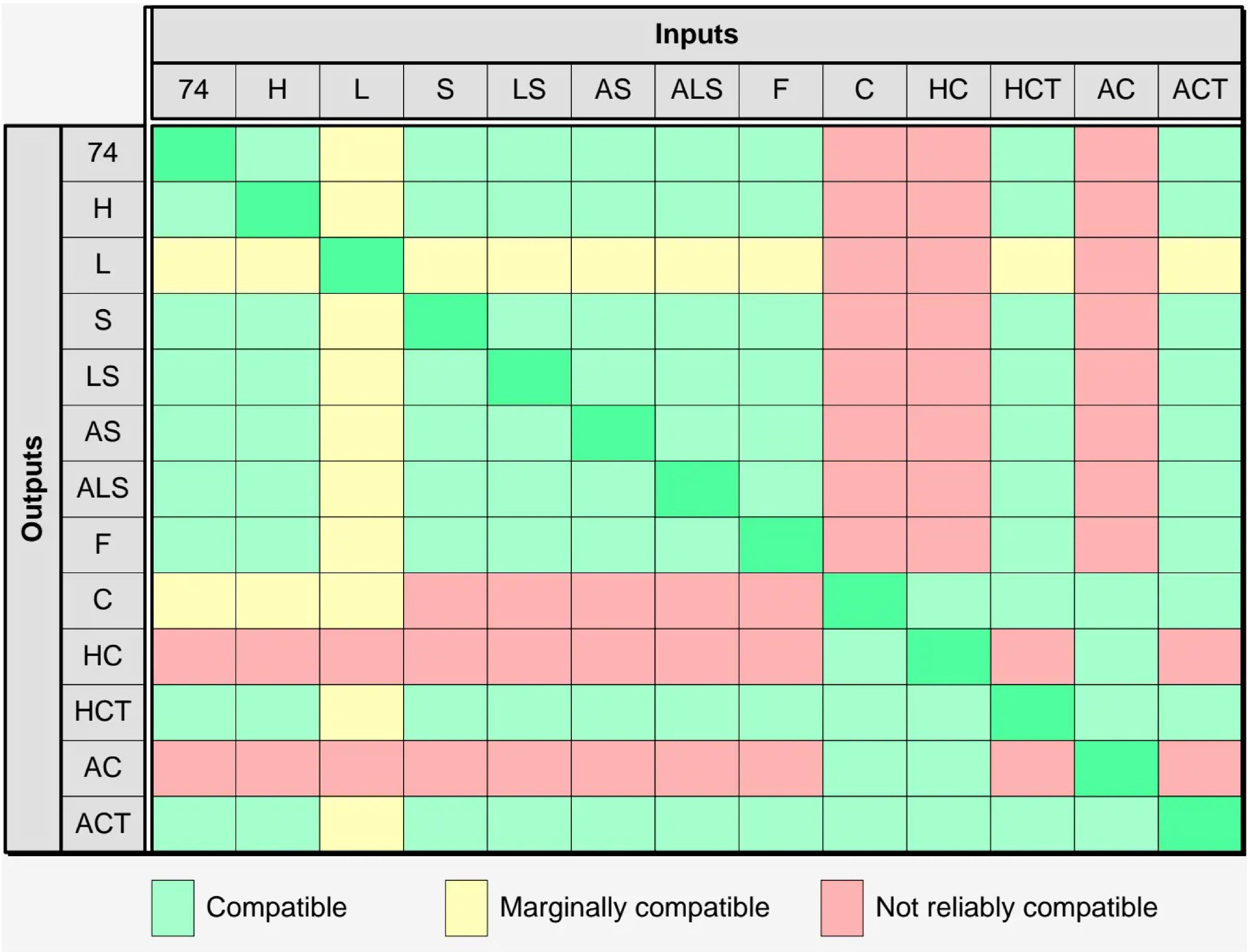
Meanwhile, CMOS stands for complementary metal-oxide-semiconductor, which employs complementary pairs of metal-oxide-semiconductor field-effect transistors (Mosfets). As they were introduced, the newer families didn't displace existing families. All the subfamilies ultimately coexisted alongside each other, which was both a blessing and a curse.

Each family provides different trade-offs between speed and power consumption. Furthermore, each family uses different transistor technologies, and each has its own input current requirements and output drive capabilities.

The question before us is which outputs can be used to drive which inputs. This is one of the classic 74xx-series rabbit holes, and it's very easy to get confused.

There are so many combinations of outputs and inputs that even professional logic designers lose track.

What we essentially need is a driver/receiver compatibility matrix depicting which 74xx-series logic families can reliably drive the inputs of other 74xx00-series logic families. Assuming we are powering all the



		Inputs												
		74	H	L	S	LS	AS	ALS	F	C	HC	HCT	AC	ACT
Outputs	74	🟩	🟩	🟨	🟩	🟩	🟩	🟩	🟩	🟥	🟥	🟩	🟥	🟩
	H	🟩	🟩	🟨	🟩	🟩	🟩	🟩	🟩	🟥	🟥	🟩	🟥	🟩
	L	🟨	🟨	🟩	🟨	🟨	🟨	🟨	🟨	🟥	🟥	🟨	🟥	🟨
	S	🟩	🟩	🟨	🟩	🟩	🟩	🟩	🟩	🟥	🟥	🟩	🟥	🟩
	LS	🟩	🟩	🟨	🟩	🟩	🟩	🟩	🟩	🟥	🟥	🟩	🟥	🟩
	AS	🟩	🟩	🟨	🟩	🟩	🟩	🟩	🟩	🟥	🟥	🟩	🟥	🟩
	ALS	🟩	🟩	🟨	🟩	🟩	🟩	🟩	🟩	🟥	🟥	🟩	🟥	🟩
	F	🟩	🟩	🟨	🟩	🟩	🟩	🟩	🟩	🟥	🟥	🟩	🟥	🟩
	C	🟨	🟨	🟨	🟥	🟥	🟥	🟥	🟥	🟩	🟩	🟩	🟩	🟩
	HC	🟥	🟥	🟥	🟥	🟥	🟥	🟥	🟥	🟩	🟩	🟥	🟩	🟥
	HCT	🟩	🟩	🟨	🟩	🟩	🟩	🟩	🟩	🟩	🟩	🟩	🟩	🟩
	AC	🟥	🟥	🟥	🟥	🟥	🟥	🟥	🟥	🟩	🟩	🟥	🟩	🟥
	ACT	🟩	🟩	🟨	🟩	🟩	🟩	🟩	🟩	🟩	🟩	🟩	🟩	🟩

🟩 Compatible
🟨 Marginally compatible
🟥 Not reliably compatible

Fig.3: the 74xx-series logic IC compatibility matrix (this assumes that everything is running at 5V).

devices at 5V, a matrix of this type is shown in Fig.3.

In some cases, the different families are logic-level compatible but not drive-strength compatible. For example, each family can typically drive 10 or more loads (devices) in the same family, but this may fall to only 1 or 2 loads when driving other families.

Suffice it to say that, although in practice we can use LS outputs to drive HC inputs (that's what we've been doing without any difficulties), this is not a reliably compatible combination. It isn't recommended, and professional engineers wouldn't do so in real-world products.

I just had a chat with a friend about this. He has drawers of 74xx-series ICs in his workshop that he's scavenged, salvaged and reclaimed over the years. He started by saying that when he's throwing a quick prototype together, he typically just looks for the correct chip type number and doesn't concern himself with the family.

During our conversation, however, he started to remember some projects that he had never been able to get working for reasons he couldn't determine, and he ended up saying, "I wonder if I was experiencing compatibility issues as depicted in your matrix."

The simplest and safest approach is to pick a single technology and stick to it. Thus, if we were starting from scratch, I'd recommend going with a kit of 74HC parts that contains all the devices we've been using in our experiments: that is, the 04 (hex inverter), the 08 (quad two-input AND), the 86 (quad two-input XOR) and the 595 (8-bit serial in, parallel out shift register).

In fact, I found just such a kit on the Temu website (**https://pemag.au/link/ac7z**).

If you want to dive deeper into all of this, a few years ago, my friend Elizabeth Simon wrote a series of *How to Read Data Sheets* columns (**https://pemag.au/link/ac80**). These cover a wide range of components, including regulators, diodes, transistors, multivibrators, relays, and, of course, logic gates and registers.

Reaction tester hardware

Returning to last month's column, having surprised myself by my lack of ability to detect a lag of less than 150~200ms between my pressing a button and observing a corresponding change in the state of an LED, I posed another problem in the form of a homework assignment for you.

As you may recall, I invited you to create a complementary Reaction Tester experiment using the orange LED on the Arduino (and the optional LED on our breadboard) along with one or more of our pushbutton switches.

The idea is for the Arduino to wait some random amount of time before illuminating the LED(s). As soon as we see the LED light up, we press the button, and the Arduino displays the time it took for us to respond on our host computer.

There are many possibilities here. For example, we could use one of our buttons to inform the program running on the Arduino when we're ready to start, and another button to respond to the LED switching on.

For no other reason than to demonstrate an alternative approach, I've decided to use a single button, which we'll call Test. We won't remove the other switches in case we need them later, but we'll pretend they don't exist (see Fig.4).

Reaction tester software

We won't go through every aspect of this program here (you can download a copy in the file named **CB-oct25-code-02.txt**). Most of our definitions remain the same as those for our perception tester, and any new definitions are readily apparent from the context.

Similarly, the **setup()** function remains largely unchanged, except we are now working with only one pushbutton switch. All we do is initialise the serial communications between our Arduino and host computer, specify which of the Arduino's pins we are using as inputs and outputs, set the outputs to their inactive levels, and clear the shift register on our breadboard.

However, it's probably worth our time to take a quick skim through the **loop()** function, as illustrated in Listing 2(a). We are using the convention that the listing number (2 in this case) corresponds to the numerical part of the matching code file name ("02" here).

On lines 72 through 75, we declare a regular **int**eger (two-byte) and three **long int**eger (four-byte) variables. On line 78, we check to see if the user has pressed the Test button. If not, we shrug our metaphorical shoulders, exit the function, then repeat the process.

When the user finally presses the button, on line 81, we switch the LED on. On line 84, we wait for the switch to stop bouncing. On line 87, we wait for the user to release the button. On line 92, we switch the LED off.

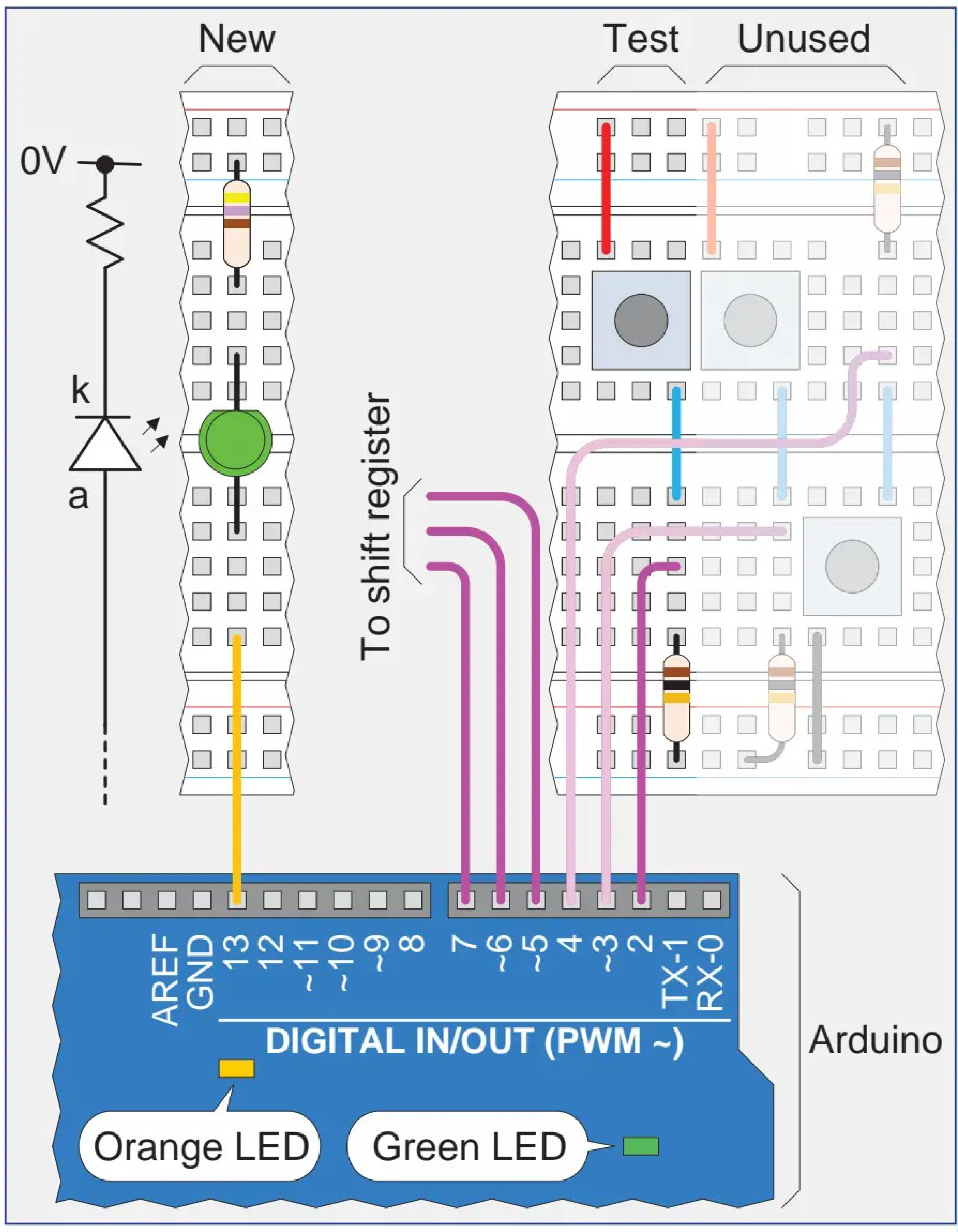


Fig.4: the breadboard layout for our reaction tester.

Listing 2(a): the main loop of the reaction tester code.

```
70  void loop ()
71  {
72    unsigned  int readySteadyGo;
73    unsigned long startTime;
74    unsigned long endTime;
75    unsigned long elapsedTime;
76
77    // Wait for user to start (test button press)
78    if ( digitalRead(PinTest) == TEST_ACTIVE)
79    {
80      // Turn LED on
81      digitalWrite(PinToLed, LED_ACTIVE);
82
83      // Wait for switch to stop bouncing
84      delay(JUST_A_BIT);
85
86      // Wait for user to release test button
87      while ( digitalRead(PinTest) == TEST_ACTIVE)
88      {
89      }
90
91      // Turn LED off
92      digitalWrite(PinToLed, LED_INACTIVE);
93
94      // Pause for some random delay
95      readySteadyGo = random(MIN_GO_DELAY, MAX_GO_DELAY);
96      delay(readySteadyGo);
97
98      // Turn LED on
99      digitalWrite(PinToLed, LED_ACTIVE);
100
101     // Time how long it takes to press test button
102     startTime = millis();
103     while ( digitalRead(PinTest) == TEST_INACTIVE)
104     {
105     }
106     endTime = millis();
107
108     // Wait for switch to stop bouncing
109     delay(JUST_A_BIT);
110
111     // Wait for user to release test button
112     while ( digitalRead(PinTest) == TEST_ACTIVE)
113     {
114     }
115
116     // Turn LED off
117     digitalWrite(PinToLed, LED_INACTIVE);
118
119     // Calculate and display the result
120     elapsedTime = endTime - startTime;
121     Serial.print("You took ");
122     Serial.print(elapsedTime);
123     Serial.println(" ms");
124   }
125 }
```

On lines 95 and 96, the program pauses for a random delay, which we've currently set to be between 5 and 10 seconds. After this delay, on line 99, we switch the LED on to indicate to the user that we're ready to proceed.

This is where things start to get interesting. The Arduino's built-in **millis()** function returns the number of milliseconds since the program started running. On line 102, we use this function to load the current time into our **startTime** variable. On line 103, we wait for the user to react to the LED lighting by pressing the button.

As soon as the user presses the button, on line 106, we record the new current time into our **endTime** variable. On line 109, we pause to ensure the switch has stopped bouncing (this isn't really necessary, but 'better safe than sorry'). On line 117, we switch the LED off again.

On line 120, we calculate the time it took the user to respond and store this in our **elapsedTime** variable. Finally, on lines 121 through 123, we display the result on the screen.

The best I achieved was around 250ms (¼ of a second). I wonder how I would have fared in my younger years. If I get my time machine working…

Before we boogie

It just struck me that we've been merrily using logic gates without noting some important portions of their backstory, so let's briefly summarise things now.

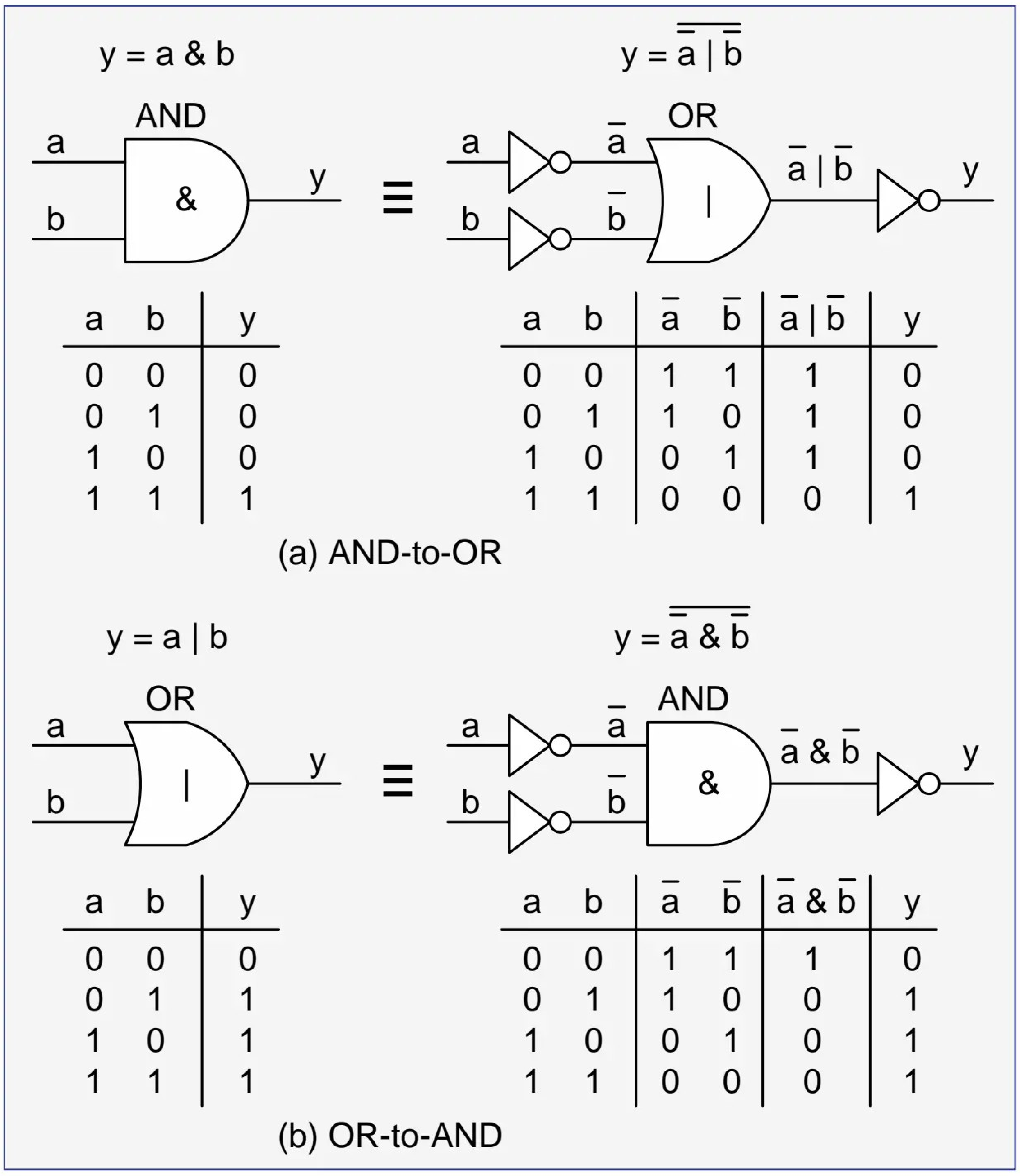
Around the 1850s, British mathematician George Boole developed a new form of mathematics, which eventually became known as Boolean algebra. Boole's intention was to use mathematical techniques to represent and rigorously test logical and philosophical arguments.

He came up with the concepts of negation, conjunction and disjunction, which we (by which I mean electronic and computer buffs) think of as NOT, AND and OR, respectively.

Sadly, Boole's work found little application outside the school of symbolic logic until the late 1930s, when, as I wrote in my book, *Bebop to the Boolean Boogie*:

> *…a graduate student at MIT, Claude Elwood Shannon, submitted a master's thesis that revolutionized electronics. In this thesis, Shannon showed that Boolean algebra offered an ideal technique for representing the logical operations of digital systems. Shannon had realized that the Boolean concepts of* ***False*** *and* ***True*** *could be mapped onto the binary digits 0 and 1, and that both could be easily implemented by means of electronic circuits.*

Logical functions can be represented using graphical symbols, equations or truth tables, and these views can be used interchangeably. Different people use different symbols. We've been using **&**, **|**, and **^** characters to represent AND, OR and XOR, because this matches the C/C++ programming language convention (which was copied by many other programming languages).


$y = a \,\&\, b$
AND
a
b
&
y
≡
$y = \overline{\overline{a} \,|\, \overline{b}}$
OR
a
b
$\overline{a}$
$\overline{b}$
|
$\overline{a} \,|\, \overline{b}$
y
$y = a \,|\, b$
OR
a
b
|
y
≡
$y = \overline{\overline{a} \,\&\, \overline{b}}$
AND
a
b
$\overline{a}$
$\overline{b}$
&
$\overline{a} \,\&\, \overline{b}$
y

a	b	y
0	0	0
0	1	0
1	0	0
1	1	1

a	b	$\overline{a}$	$\overline{b}$	$\overline{a} \,\|\, \overline{b}$	y
0	0	1	1	1	0
0	1	1	0	1	0
1	0	0	1	1	0
1	1	0	0	0	1

(a) AND-to-OR

a	b	y
0	0	0
0	1	1
1	0	1
1	1	1

a	b	$\overline{a}$	$\overline{b}$	$\overline{a} \,\&\, \overline{b}$	y
0	0	1	1	1	0
0	1	1	0	0	1
1	0	0	1	0	1
1	1	0	0	0	1

(b) OR-to-AND

Fig.5: some of the most important DeMorgan transformations.

It's also common to use the **!** character to represent NOT (also a C/C++ convention). However, another popular alternative is to use over-bars (horizontal lines placed over a character or group of characters) to indicate negation (similar to how we've been using over-bars to indicate active-low signals, which are, in a sense, a negated form of logic).

Feeling transformed

Another British mathematician, Augustus DeMorgan, who was a contemporary of Boole's, came up with a set of rules called DeMorgan transformations. These have all sorts of uses. For our purposes here, we need to note only that they can be used to transform AND operations into ORs and vice versa.

Let's start by transforming an AND into an OR, illustrated in Fig.5(a). Consider this in the form of the equation $y = a \,\&\, b$. We commence by inverting the inputs **a** and **b** to give $\overline{a}$ and $\overline{b}$. Next, we replace the **&** with an **|** to give $\overline{a} \,|\, \overline{b}$. Finally, we invert the entire function to give $y = \overline{\overline{a} \,|\, \overline{b}}$.

Now, consider implementing the same functionality following the same sequence using logic gates. In this case, we start by inverting the inputs using NOT gates. Next, we swap the AND gate for an OR gate. Finally, we invert the output from the whole function by attaching a NOT to the output from the OR gate.

I've no idea
why they
call me
"Max Max"!

If you examine the truth tables for the initial AND operation, along with its DeMorgan counterpart, as depicted in Fig.5(a), you'll see that this works a treat. We can use a similar approach to transform an OR into an AND, as illustrated in Fig.5(b).

Now, your initial reaction might be that this is horrendously complicated, but it's really not as bad as it seems. In the case of the AND-to-OR shown in Fig.5(a), for example, at first glance it looks like we've transformed a single level of logic (the AND) into three levels of logic (the NOTs on the inputs, the OR, and the NOT on the output).

Does this mean that our new OR will switch slower than our AND? Well, let's think about this a bit. In last month's column, we noted that, 'under the bonnet', an AND is actually formed from a NAND followed by a NOT, which we could write as NAND → NOT, which means our AND already contains two levels of logic.

In the case of our transformed function, we could think of this as NOT → OR → NOT. In reality, however, our OR is actually formed from a NOR followed by a NOT, which means our transformed function is actually NOT → NOR → NOT → NOT, which is four levels of logic (eek).

But, from last month's column, we also know that two NOTs in series (ie, NOT → NOT) perform the same job as not having any NOTs at all, which means we can remove the two NOTs from the output of the NOR. All this boils down to the fact that we've changed a NAND → NOT path into a NOT → NOR equivalent, both of which have two levels of logic.

There's more to it, but I think this is enough to keep us cogitating for a while.

Why am I waffling on about all this? Well, I'm a digital logic designer by trade. We used to do this sort of thing by hand all the time in the days of yore when I was bright-eyed and bushy-tailed.

These days, when developing new digital integrated circuits, logic designers have advanced tools that do all the heavy lifting for them; however, it's still helpful to know how to do this when creating breadboard prototypes and working with a limited supply of logic gate options.

The way I think about this is that, when solving mathematical problems, it's great to have a calculator, but it's also handy to know how to perform the calculations by hand in case our batteries die. The same thing applies to logic. It's great to have clever tools, but it's also satisfying to understand how they work their magic.

Would you like to see a DeMorgan transformation being employed to solve a real-world problem? I was sure you would say yes!

DeMorgan in action

In our July 2025 *Cool Beans* column, we introduced the concept of linear feedback shift registers (LFSRs) as a method of generating pseudorandom numbers (not truly random numbers, but close enough for many jobs).

As part of this, we decided that we needed a two-input XNOR gate. The problem was that we only had a quad two-input XOR gate in our kit of parts.

Our initial solution was to invert the output of our XOR with a NOT gate to provide the required XNOR functionality (see Fig.6).

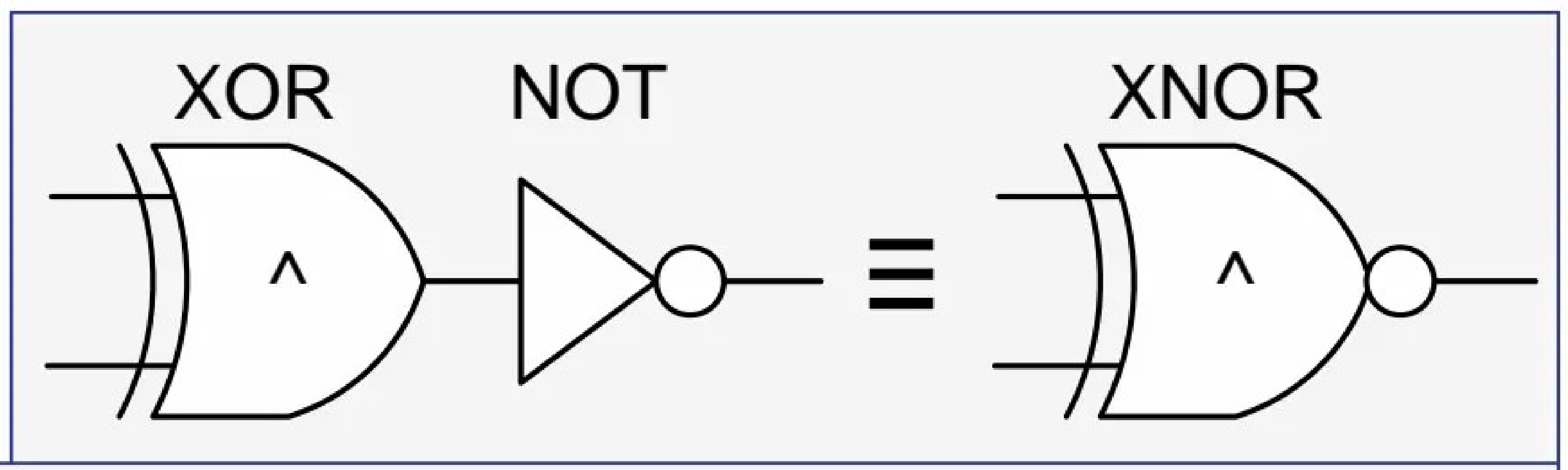


Fig.6: creating an XNOR gate from XOR and NOT gates.

Also in that issue, I posed two homework tasks. The first was to suppose we had only our 74xx86 chip containing its four two-input XOR gates (so, no 74xx04 with its six NOT gates and no 74xx08 with its four two-input AND gates). Your job was to implement an XNOR function using only XOR gates. Last month, we presented one possible solution, depicted in Fig.7.

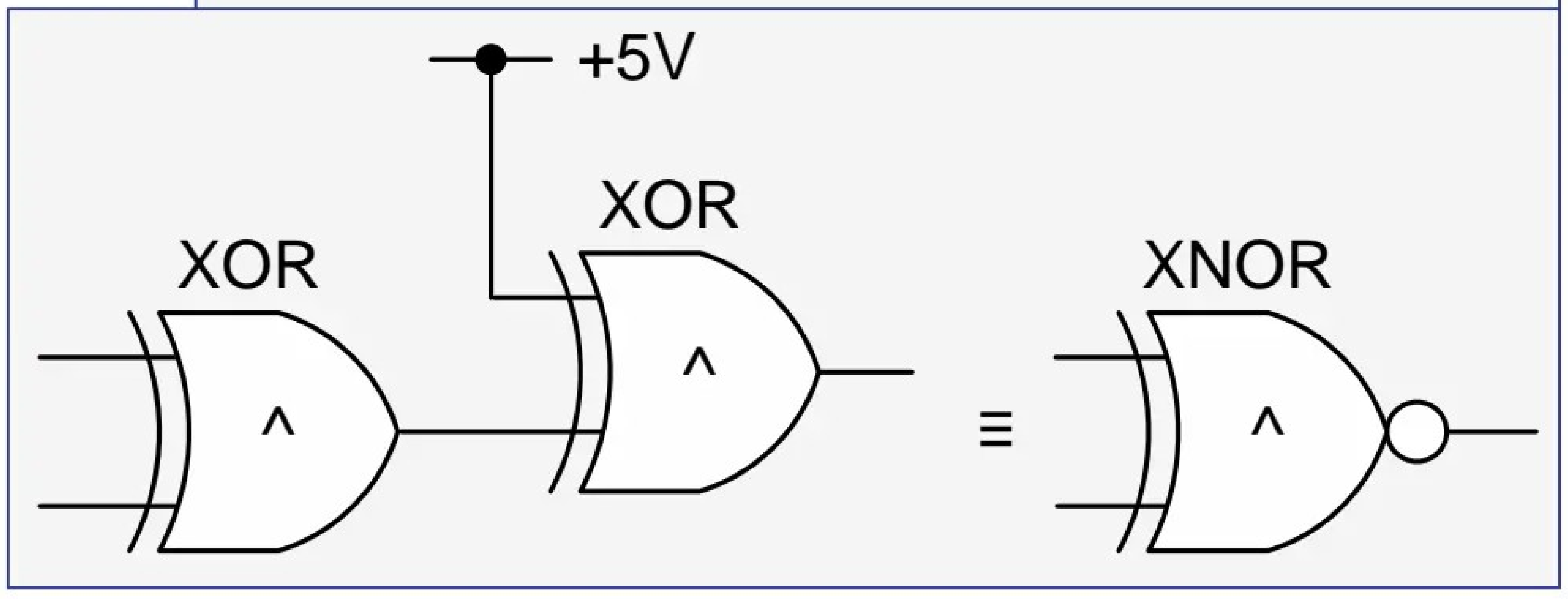
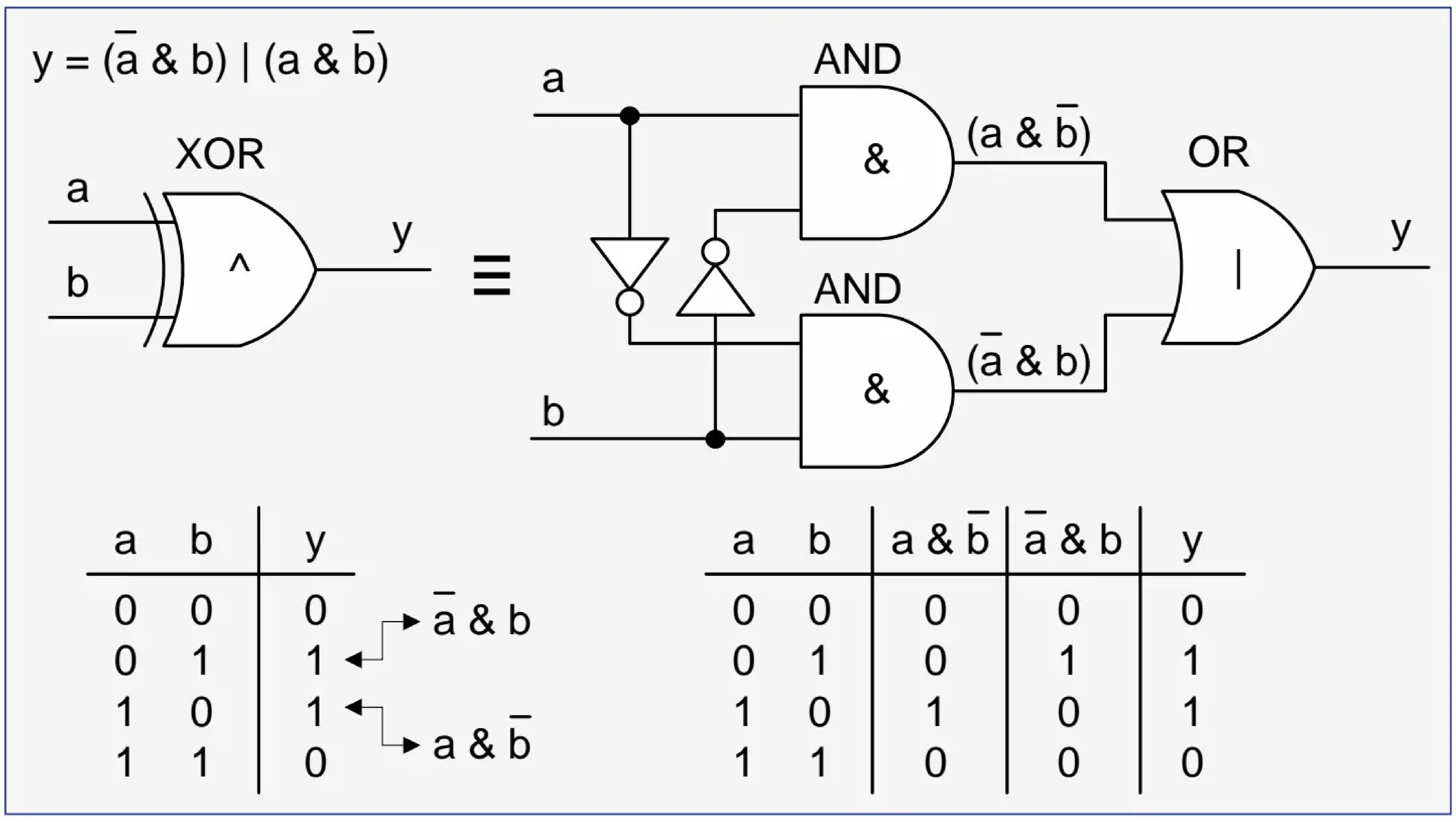


Fig.7: creating an XNOR gate from two XOR gates.

The second homework task was to suppose we had only our 74xx04 with its six NOT gates and our 74xx08 with its four two-input AND gates (so, no 74xx86 with its four two-input XOR gates), and to implement the desired two-input XNOR function using only AND and NOT gates

Once again, we presented one possible solution last month. We started by implementing an XOR using four NAND gates, which were themselves implemented as ANDs followed by NOTs, and then we transformed this XOR into an XNOR by removing the NOT from the final output signal.

Now that we are familiar with DeMorgan transformations, we can devise other methods to create our XNOR. In line with our earlier solutions, let's start by implementing an XOR, as illustrated in Fig.8.



a	b	y
0	0	0
0	1	1
1	0	1
1	1	0

a	b	$a \,\&\, \overline{b}$	$\overline{a} \,\&\, b$	y
0	0	0	0	0
0	1	0	1	1
1	0	1	0	1
1	1	0	0	0

Fig.8: creating an XOR gate from two NOT gates, two AND gates and an OR gate.

The approach we might take to do this is as follows. We start by drawing the XOR symbol and its corresponding truth table.

From the truth table, we identify the rows for which the output is 1.

In natural language, from the second row, we can state that the **y** output is 1 if the **a** input is 0 AND the **b** input is 1. From the third row, we can state that the **y** output is 1 if the **a** input is 1 AND the **b** input is 0.

We can combine these to realise that the **y** output is 1 if the **a** input is 0 AND the **b** input is 1, OR if the **a** input is 1 AND the **b** input is 0. We can represent this sentiment as a Boolean equation, or as a combination of logic gates.

As an aside, due to the similarities between arithmetic and logical operators, the Boolean **&** (AND) operator is known as a logical multiplication, or *product*, while the Boolean **|** (OR) operator is known as a logical addition or *sum*.

Thus, the equation $y = (\overline{a} \,\&\, b) \mid (a \,\&\, \overline{b})$ would be said to be in a sum-of-products (SOP) form. There's also a product-of-sums (POS) counterpart form, but we'll leave that for another day.

The only problem is that we don't have an OR gate. Well, that's what we're telling ourselves. In fact, we do have quad two-input OR, NAND, and NOR gates in our kit of parts (74xx32, 74xx00, and 74xx02, respectively), but we are pretending that we have only AND and NOT gates for the purpose of this exercise.

Happily, we now know that we can use a DeMorgan transformation to replace the OR gate shown in Fig.8 with the AND and three NOT gates shown in Fig.5(b).

Last, but certainly not least, since we now have an XOR but we really want an XNOR, all we need to do is remove the NOT gate driving the output. I feel a little "tra-la" is in order.

How about one more?

As we've discussed, we began by creating an XNOR by inverting the output from an XOR with a NOT, and this approach has guided all our subsequent implementations. Based on our newfound knowledge, suppose we decide to come at things from a different angle; that is, starting with the truth table for our desired

XNOR function, as illustrated in Fig.9.

Once again, we can use a DeMorgan transformation to replace the OR gate shown in Fig.9 with the AND and three NOT gates shown in Fig.5(b). In this case, however, we don't need to subsequently remove the final NOT, because we already have the XNOR function we crave.

Universal gates

As a point of interest, two-input NAND gates and two-input NOR gates are known as "universal gates" because you can use either of them to create any of the other logic gate types. In the case of the NAND, for example, we can form a NOT by tying the two inputs of a NAND together, as illustrated in Fig.10.

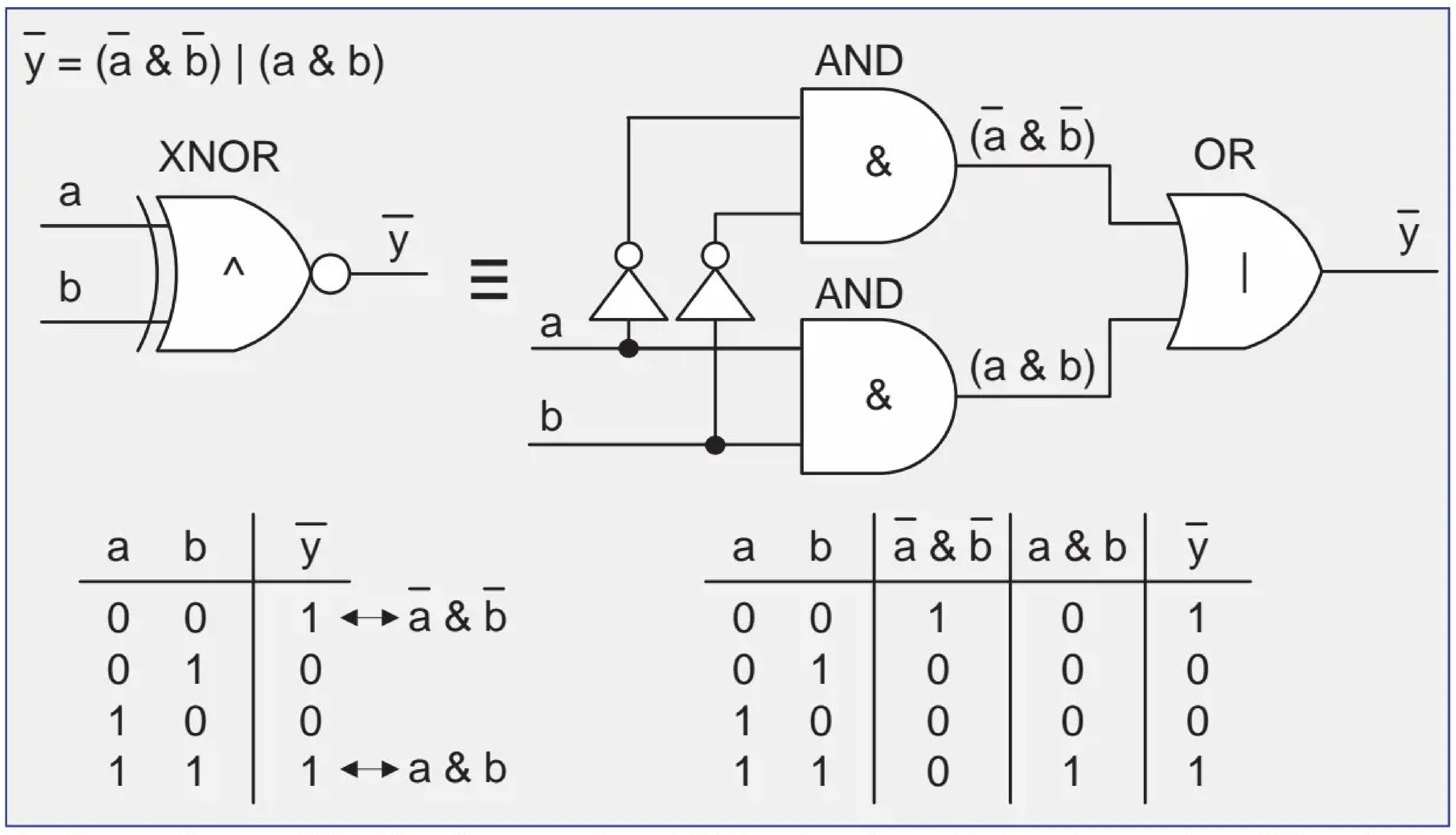
We already know that we can form an AND by inverting the output of a NAND with a NOT. Now we know that we can form an AND using two NANDs. Similarly, we can transform an AND into an OR, and… so it goes.

In fact, any combinational logic circuit, no matter how complex, can be represented as a network of NAND (or NOR) gates. Logic synthesis tools for designing digital ASICs (application-specific integrated circuits) often perform such transformations internally for optimisation and mapping!

Next time

I can't believe it. I've allowed you to sidetrack me once again (shame on you!). But I won't let you divert me anymore. In our next column, we will definitely return to building our retro games console, wiring up the 7-segment displays and the eight push-button switches on the front panel, or my name's not Max the Magnificent.

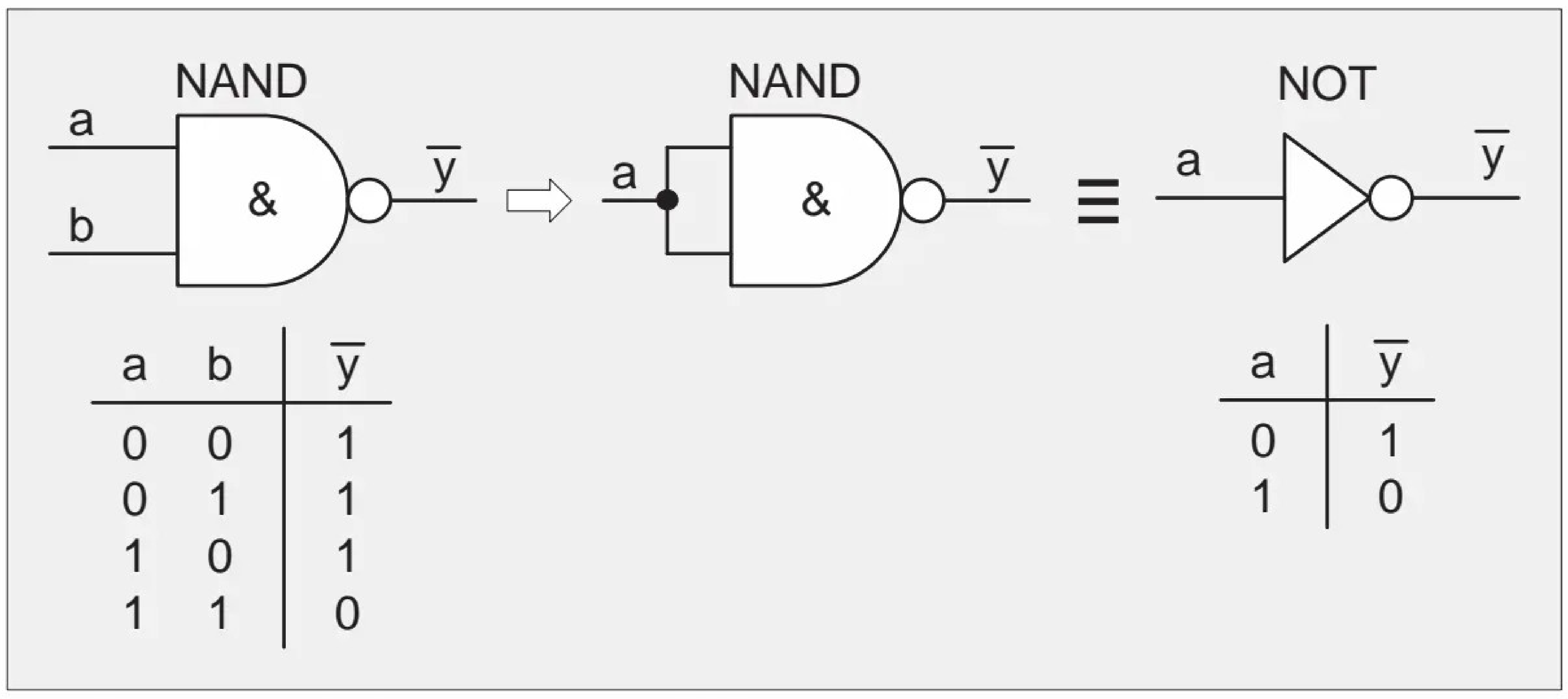
As always, if you have any thoughts you'd care to share on anything you've read here, please feel free to email me at max@clivemaxfield.com. PE


$\overline{y} = (\overline{a} \& \overline{b}) | (a \& b)$
XNOR
a
b
^
$\overline{y}$
≡
AND
&
$(\overline{a} \& \overline{b})$
OR
|
$\overline{y}$
AND
&
(a & b)
a
b

a	b	$\overline{y}$	
0	0	1	↔ $\overline{a} \& \overline{b}$
0	1	0	
1	0	0	
1	1	1	↔ a & b

a	b	$\overline{a} \& \overline{b}$	a & b	$\overline{y}$
0	0	1	0	1
0	1	0	0	0
1	0	0	0	0
1	1	0	1	1

Fig.9: creating an XNOR gate using two NOT gates, two AND gates and an OR gate.



a	b	$\overline{y}$
0	0	1
0	1	1
1	0	1
1	1	0

a	$\overline{y}$
0	1
1	0

Fig.10: using a NAND gate to form a NOT gate.

Devices and components used in this issue

Arduino Uno R3 microcontroller module	https://pemag.au/link/ac2g
Solderless breadboard	https://amzn.to/3O2L3e8
4-inch (10cm) jumper wires (optional)	https://pemag.au/link/ac2l
8-inch (20cm) jumper wires (male-to-male)	https://amzn.to/3O4hnxk
LEDs (assorted colours)	https://amzn.to/3E7VAQE
Resistors (assorted values)	https://amzn.to/3O4RvBt
Ceramic capacitors (assorted values)	https://pemag.au/link/ac2i
16V 100µF electrolytic capacitors	https://pemag.au/link/ac2j
Momentary pushbutton switches	https://amzn.to/3Tk7Q87
8-Segment DIP red LED bar displays	https://pemag.au/link/ac2c
Kit of popular SN74LSxx chips*	https://pemag.au/link/ac2k
74HC595 8-bit shift registers*	https://pemag.au/link/ac1n
Kit of popular SN74HCxx chips**	https://pemag.au/link/ac7z

** This kit provides all the ICs we need, instead of the two * entries.

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The Fox Report

Barry Fox's technology column

Christmas comes early this year

Christmas for the IT trade comes in October this year. That's when Microsoft stops sending out free security patches for Windows 10 PCs. The hundreds of millions of computers that cannot be upgraded from Windows 10 to 11 will immediately be at increased risk from malware.

The only safe way to continuing using a Windows 10 PC, other than paying Microsoft or a third-party for an extended update service will be to completely disconnect it from the Internet.

We've previously detailed some workarounds, which let perfectly good PCs that fall short of Microsoft's draconian requirements sneak an upgrade from 10 to 11. So far, none of these has been usable by 'average' users, because they involve registry editing. The latest trick, called FlyBy, is easier to use, but still not for novices (more on this next month).

Faced by mounting bad publicity, Microsoft has offered some sops that mostly involve paying money or credits for the most essential updates. It remains to be seen whether Microsoft will climb down further.

In the meantime, many people will do as Microsoft advises and just buy a new PC.

I reckoned the time was right to buy a new laptop to replace an old and painfully slow machine. It should have been a happy experience; it wasn't, but I learned a lot of lessons that I'll now pass on.

Buying by mail order did not appeal. I wanted to look at screens and feel keyboards, so I took a train to London to browse the on-display stock in Currys emporium in Tottenham Court Road.

I'll gloss over the generally lackadaisical lack of interest shown by the sales staff, and just say I ended up choosing an MSI Prestige 14 Evo laptop, advertised online and instore as costing £549.

Within a week or so, the price had been reduced by £50, but that was just my tough luck, so I swallowed it. What I didn't swallow was Currys' pricing tactic, and the way my query on it was handled.

Like many big stores, Currys routinely (and legitimately) hard-sells 'extras' (eg, computer software and insurance services) at the point of sale. It's where the big profits lie – like popcorn in a cinema. I have learned to say "no thanks" to everything extra unless it is free.

In this case, the in-store salesman offered, and I accepted, a complimentary copy of McAfee anti-virus. I paid in full by credit card, and the receipt showed £549 for the Evo laptop and £11.99 for McAfee, with this written off by a Promotional Offer of £12.

So far, so good. But I had to ask the salesman for an invoice for tax purposes. Instead of printing it for me in-store, he emailed it to me. So I could not check it in-store.

When I later had home access to email and a printer, I discovered that the invoice itemised the sale as £537 for the laptop, plus another £11.99 for a mysterious extra item, cryptically identified only as "MCA Total Pro.T.1DEV". I emailed back to the salesman, querying this. The replies I got were largely incomprehensible

For your entertainment, I quote: "The McAfee was giving to you but would be of discounted on the system the invoice would just have the sale but on the receipt we gave you you'll see that it was discounted for you as an complimentary item… instead of just paying for the laptop for the same price he gave you the McAfee as a bundle."

Eventually, it became clear that the mysterious extra was a copy of McAfee. So Currys was offering a complimentary extra (£11.99 McAfee with a £549 laptop), overtly adding the £11.99 cost to the bill, cancelling it with a "Promotional Offer -£12", but all the time covertly burying the cost of the same extra in the advertised cash price for the laptop.

So, the customer unwittingly pays for an extra they think they are getting for free.

The salesman finally offered me a refund of £12, but stipulated that I would have "to come down to our store for a manual refund with your card". In other words, spend more than my loss to take a train to London to recoup my loss.

Over many years, I have bought many devices from Currys with no problems. So I wrote (twice) to Currys CEO Alex Baldock asking if he was aware of how his sales staff are behaving. I heard nothing back.

Currys has now stopped offering customers the chance to contact or complain by email or web form, with emailed acknowledgement or receipt. So I had to snailmail my letters to the CEO and snailmail my complaint to Currys Complaints, PO Box 1687, Sheffield S2 5YA. I also contacted Currys Press Office.

It took a month to hear back. The Press Office parroted the standard line "we will not be providing a comment at this time". A complaints clerk phoned, saying that the CEO had not received my letter. No surprises there then. She seemed generally baffled by the pricing tangle and promised to bone up and get back to me.

She was "unable" to give any useful contact information in case I needed to chase her. And surprise, surprise, she didn't get back to me.

Is it really possible that the Evo laptop has been the only product on sale that came with a complimentary extra that was actually charged for in the product price? I raised the question with the Advertising Standards Authority, who told me to go to the Citizens Advice Service, who might then take it up with the Trading Standards Office.

I did just that. But in the meantime, I followed the golden advice I now give to anyone who feels they have been stitched up. I contacted my bank and asked for a £12 chargeback refund. Significantly, the bank looked at the evidence I provided and immediately refunded the over-payment.

Significantly, Currys, made no attempt at challenging the refund. The usual routine, I am told by insiders, is for the dealer to make at least a token rebuttal.

Trading Standards held that "there is the potential for 'confusion' at the way the invoices after the sale are presented which Currys need to be made aware of... will discuss this with Currys and suggest how they could change invoicing in the future to avoid confusion (and) also express concern regarding the customer service issues that you have raised to us."

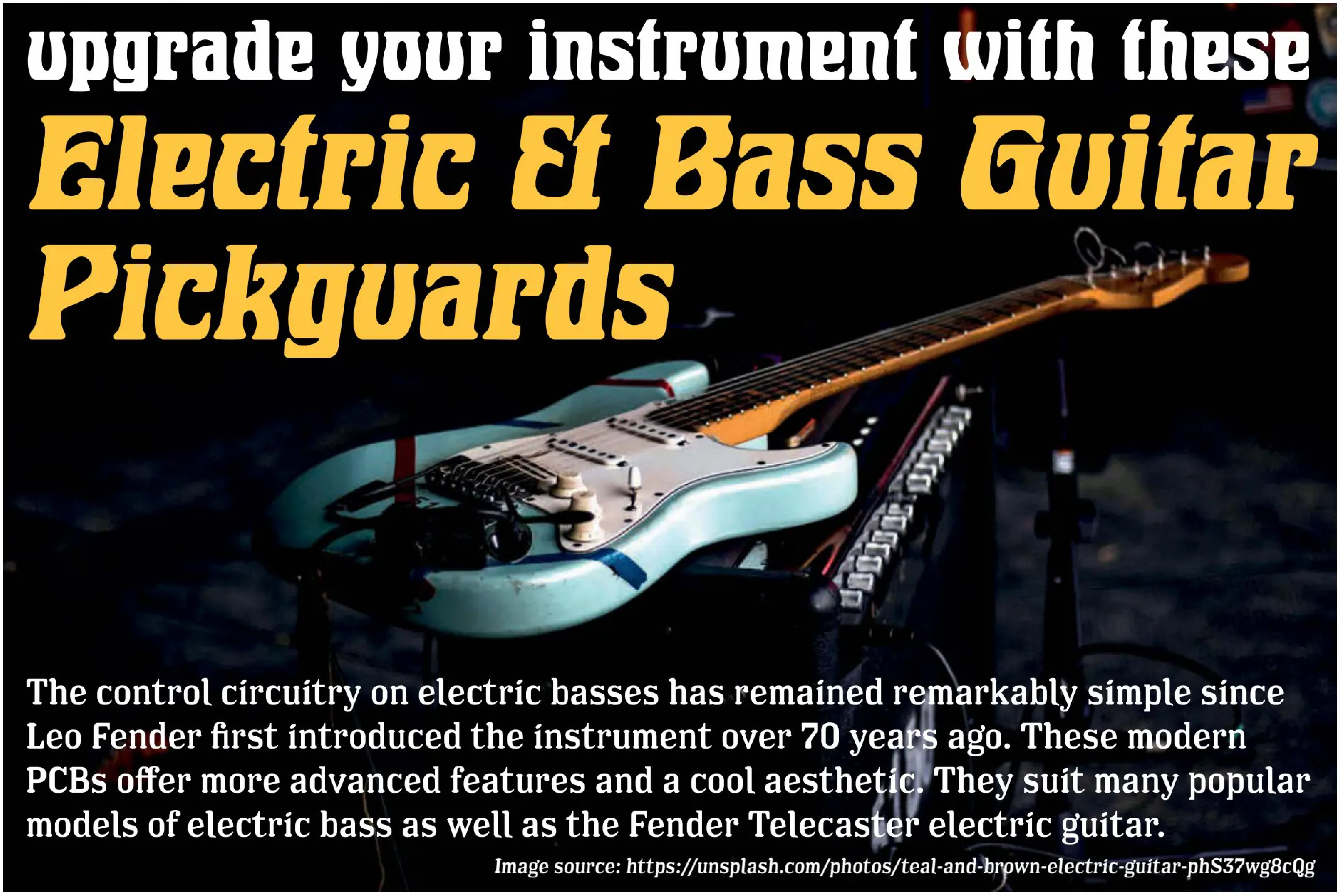
After a lot more of my time-consuming persistence, Currys complaints finally got back to me, sort-of apologised, made me an ex-gratia payment, which was a pinprick in my time wasted, and confirmed that the CEO does not reply to customers' letters.

So, if you feel done down in a deal, document your case and ask your bank to give you a chargeback refund. If your bank thinks you are in the right, the dealer is stuck with the problem of proving they didn't do you down. That's what matters. Shift the burden of proof from the customer to the dealer.

Continued on page 51...

Project by Brandon Speedie

upgrade your instrument with these Electric & Bass Guitar Pickguards



The control circuitry on electric basses has remained remarkably simple since Leo Fender first introduced the instrument over 70 years ago. These modern PCBs offer more advanced features and a cool aesthetic. They suit many popular models of electric bass as well as the Fender Telecaster electric guitar.

Image source: https://unsplash.com/photos/teal-and-brown-electric-guitar-phS37wg8cQg

Most electric guitars and basses have circuitry built into the instrument's body. Typically, it includes a passive network of potentiometers and capacitors to give the musician control of output volume and 'tone'. The standard configuration is shown in Fig.1.

The volume control is a potentiometer that divides the audio signal voltage from the pickup(s). The tone control is an adjustable low-pass filter to reduce the amount of treble and therefore change the instrument's sound to suit different music and playing styles.

Its simplicity has made it a popular circuit, remaining relatively unchanged since the early 20th century. However, it has some weaknesses. These updated circuits aim to correct some of those shortcomings and add some handy new features.

Most electric instruments, including guitars and basses, use passive inductive pickups to sense the vibrations of the metal strings, converting them into electrical signals that can be amplified.

Pickups

There are predominantly two types of electromagnetic pickups used on guitars and basses. The most common is the 'single coil' type, so called because it is constructed of a single inductor wrapped around a set of permanent magnets. The magnets are made of an iron alloy known as alnico (aluminium, nickel & cobalt), which are positioned under the instrument strings to form 'pole pieces'.

Wrapped around these magnets are several thousand turns of enamelled copper wire (see Fig.3). The pole pieces magnetise the strings, producing a changing magnetic field for the copper coil when plucked. This movement induces a voltage in the coil, which is ultimately sent to an amplifier for playback or recording.

The other type of pickup is called a "humbucker" because it can cancel interference and therefore reduce the hum induced by noisy sources such as nearby transformers and fluorescent lighting. The humbucker has two coils mounted next to each other in the

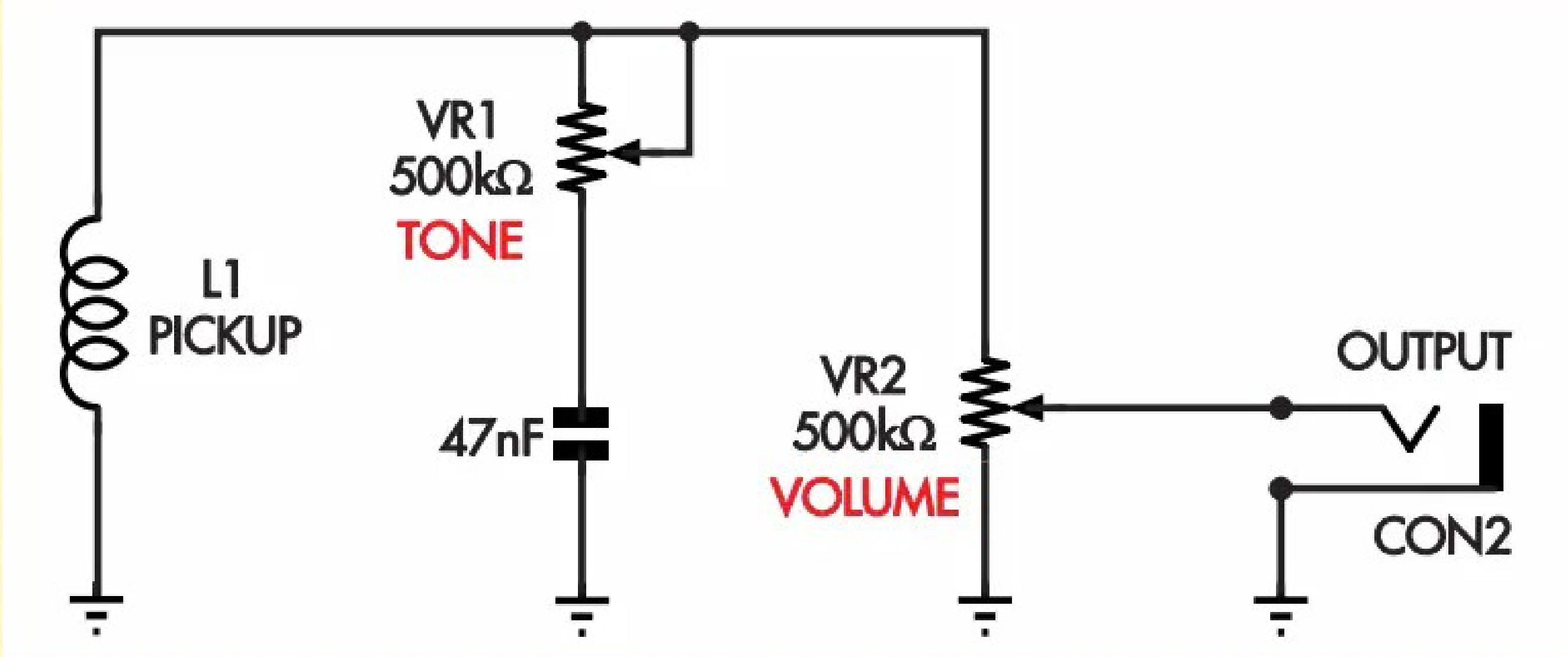


Fig.1: a typical control circuit on an electric guitar or bass. The 'tone' control is an adjustable low pass filter, while the volume control is an adjustable voltage divider.

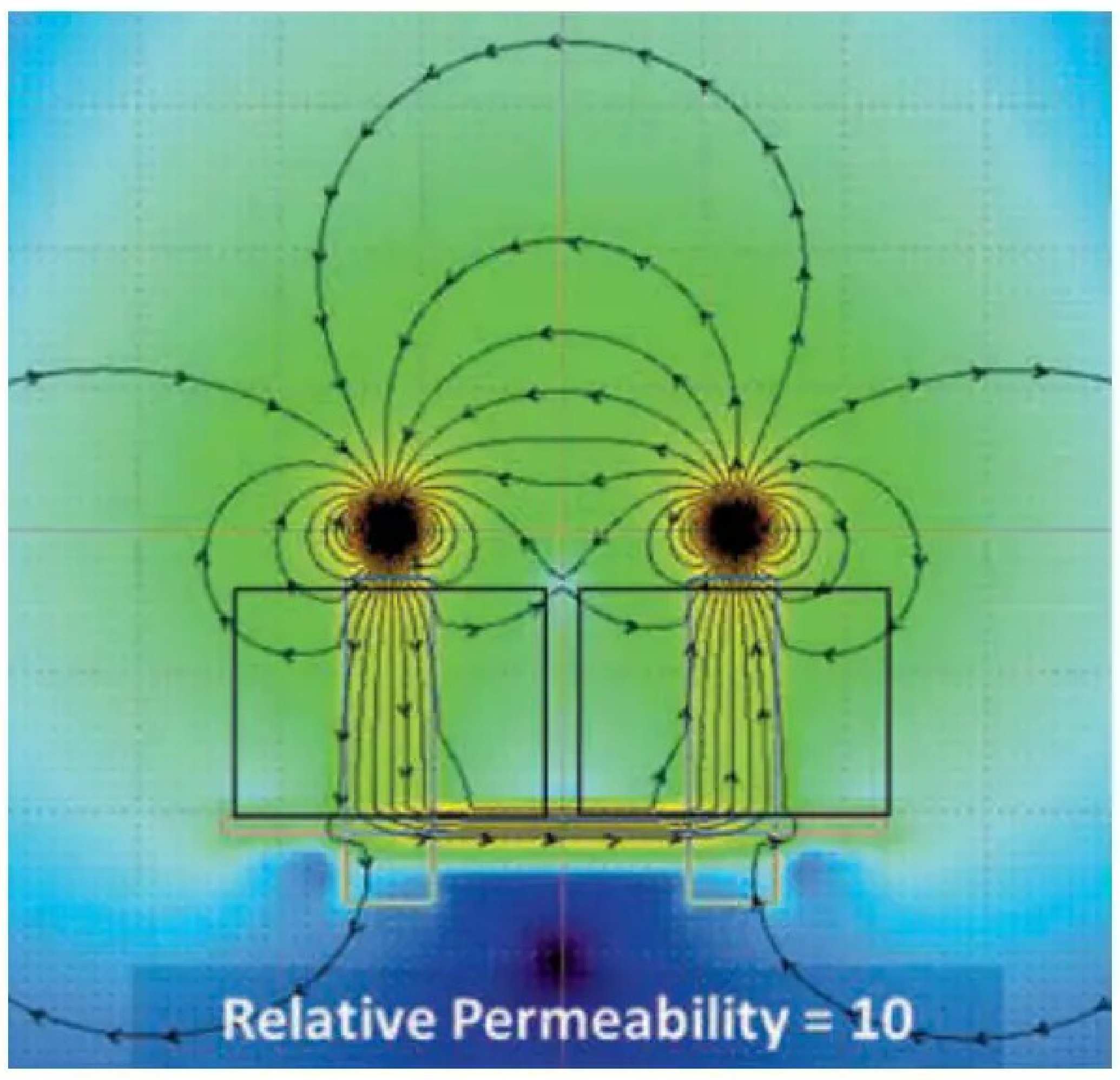
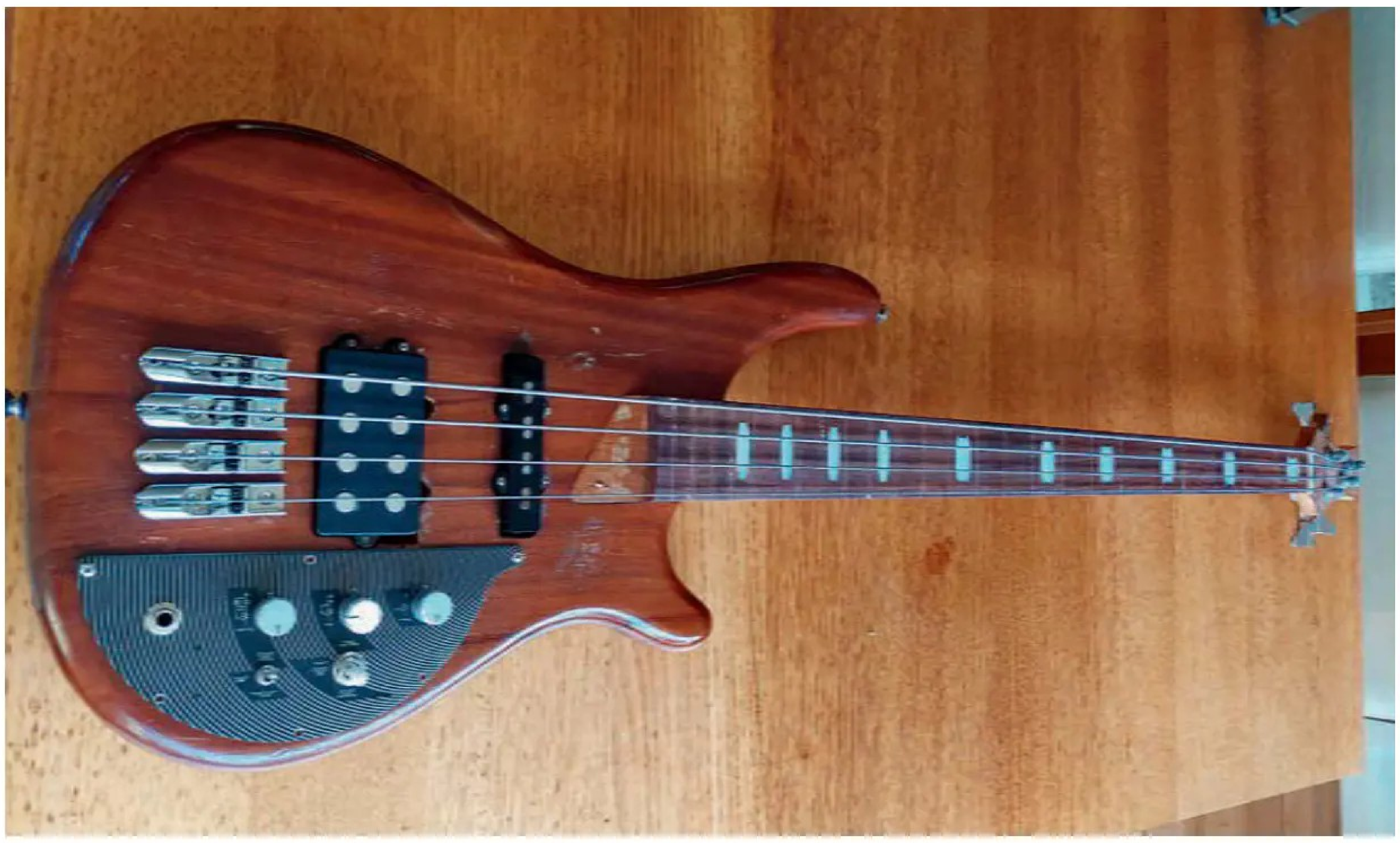
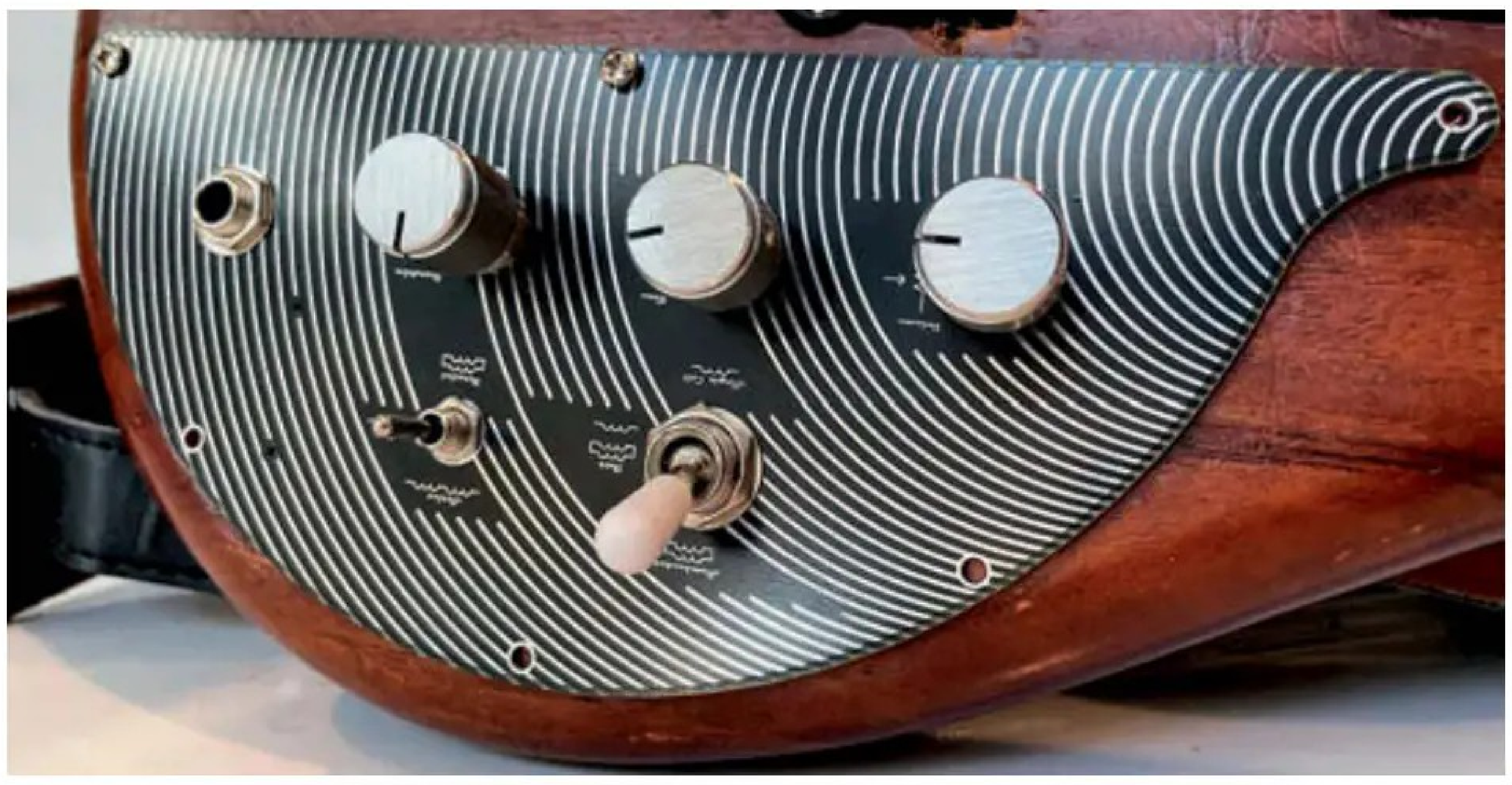


Fig.2: the magnetic field lines around a typical humbucker pickup. ***Source: Lawing Musical Products – https://pemag.au/link/abw4***





Photos 1 & 2: the outside surface of the T-style bass pickguard has a nice tinned pattern. The labels aren't upside-down; at least, not from the player's perspective!

same package. One coil has its north magnets facing the strings, while the other has its south poles facing the strings (Figs.2 & 3).

The coils are wired 180° out of phase, so any external interference that impinges on the pickup will induce an opposing voltage in each coil that is therefore cancelled out. Any (wanted) voltage induced by the strings will be out-of-phase due to the opposite orientation of the pickup magnets, and those signals will reinforce due to the out-of-phase wiring.

Humbuckers are known for their stronger and fuller sound but tend to lack clarity and brightness compared to a single coil. These differing characteristics, as well as pickup placement, can be used by the instrument designer to influence its overall voicing.

Pickups placed towards the neck of the instrument tend to have more bass and sound more mellow, while bridge-mounted pickups will sound brighter with strong mid-range and treble sounds. The electronic control circuitry allows the musician to select these different pickups and control the final sound.

I have designed four different circuits with custom-shaped PCBs to suit some of the more common or interesting electric basses and guitars.

J&D Luthiers T-Style electric bass

The inspiration for this project came during the restoration of an old bass guitar, known as a T-Style, from the Australian designer J&D Luthiers (see Photo 1). The existing circuitry was mounted directly to timber veneer, which was showing its age and needed replacement. I routed the veneer off, leaving a hole in the body to be covered by a new fascia.

The obvious choice for a new material would be custom-machined sheet metal, but making that is time consuming and quite expensive. Most parts of this nature are also chrome-coated, which adds further expense. Instead, I decided to make a new fascia from a printed circuit board (PCB), which could act as both a visually appealing fascia and house the new circuitry (see Photo 2).

The new circuit is shown in Fig.4 and Photo 3. The instrument features two pickups: a single coil near the neck and a humbucker near the bridge. These wire directly to CON1, a 7-way screw terminal.

Both coils' negative ends and the humbucker shield are grounded, along with the instrument drain wire. The drain is electrically connected to the

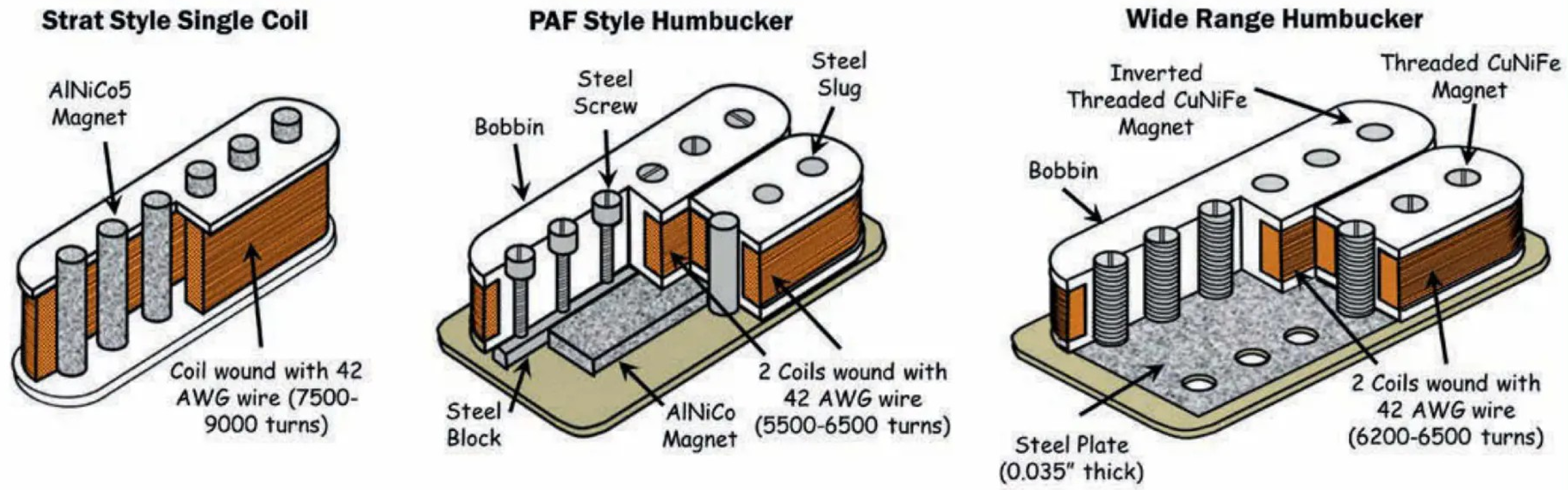

Strat Style Single Coil
AlNiCo5 Magnet
Coil wound with 42 AWG wire (7500-9000 turns)
PAF Style Humbucker
Steel Screw
Steel Slug
Bobbin
Steel Block
AlNiCo Magnet
2 Coils wound with 42 AWG wire (5500-6500 turns)
Wide Range Humbucker
Inverted Threaded CuNiFe Magnet
Threaded CuNiFe Magnet
Bobbin
Steel Plate (0.035" thick)
2 Coils wound with 42 AWG wire (6200-6500 turns)

Fig.3: the internal construction of a single coil pickup, an early "PAF" Humbucker with bottom-mounted magnet and steel pole pieces, and a more modern form of humbucker with alnico magnet pole pieces. ***Source:*** ***https://lawingmusicalproducts.com/dr-lawings-blog/the-wide-range-humbucker-and-the-genius-of-seth-lover***

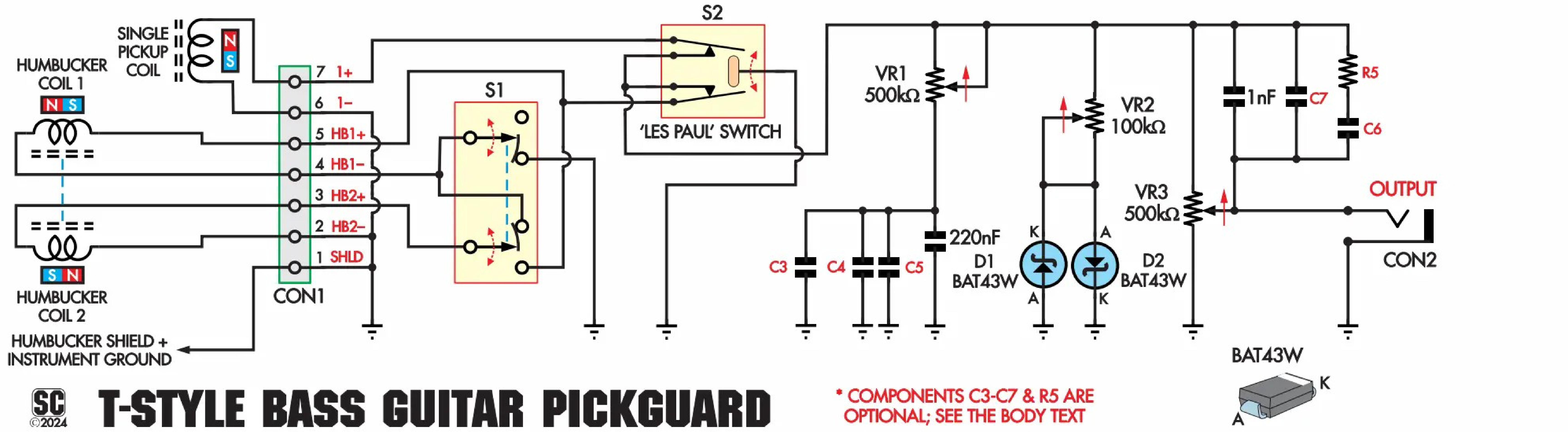


Fig.4: my new circuit for the J&D Luthiers T-Style bass allows you to select which pickups are active and the configuration of the humbucker. It also provides tone, overdrive and volume controls and has compensation so that the frequency response doesn't change too much with volume level.

bridge, to suppress interference induced on the strings and metal hardware.

The 'middle' of the humbucker (coil one negative, coil two positive) connects to switch S1, a DP3T toggle switch with a slightly unusual on/on/on switching pattern, as shown in Fig.5. This provides the option of series/split/parallel selection for the humbucker coils.

When in the split position (centre), one of the coils is grounded, so the humbucker operates as a single coil, giving a clear and bright tone. When in the series position (down), the coils are in series. This gives the strongest output and a rich tone but less brightness than a single coil.

When in the parallel position (up), the coil one negative is grounded and the coil two positive connects to the output, placing the coils in parallel. This gives a tone somewhere between the other two modes.

The output of the humbucker switch leads to S2, the bridge/neck pickup selector switch. In the bottom position, the neck single coil will be active; in the top position, the bridge humbucker will be selected (in whatever mode S1 has it operating in); and, in the centre position, both pickups are active.

Switches S1 & S2 provide a lot of flexibility for the musician, selecting between a total of seven different configurations for the two pickups.

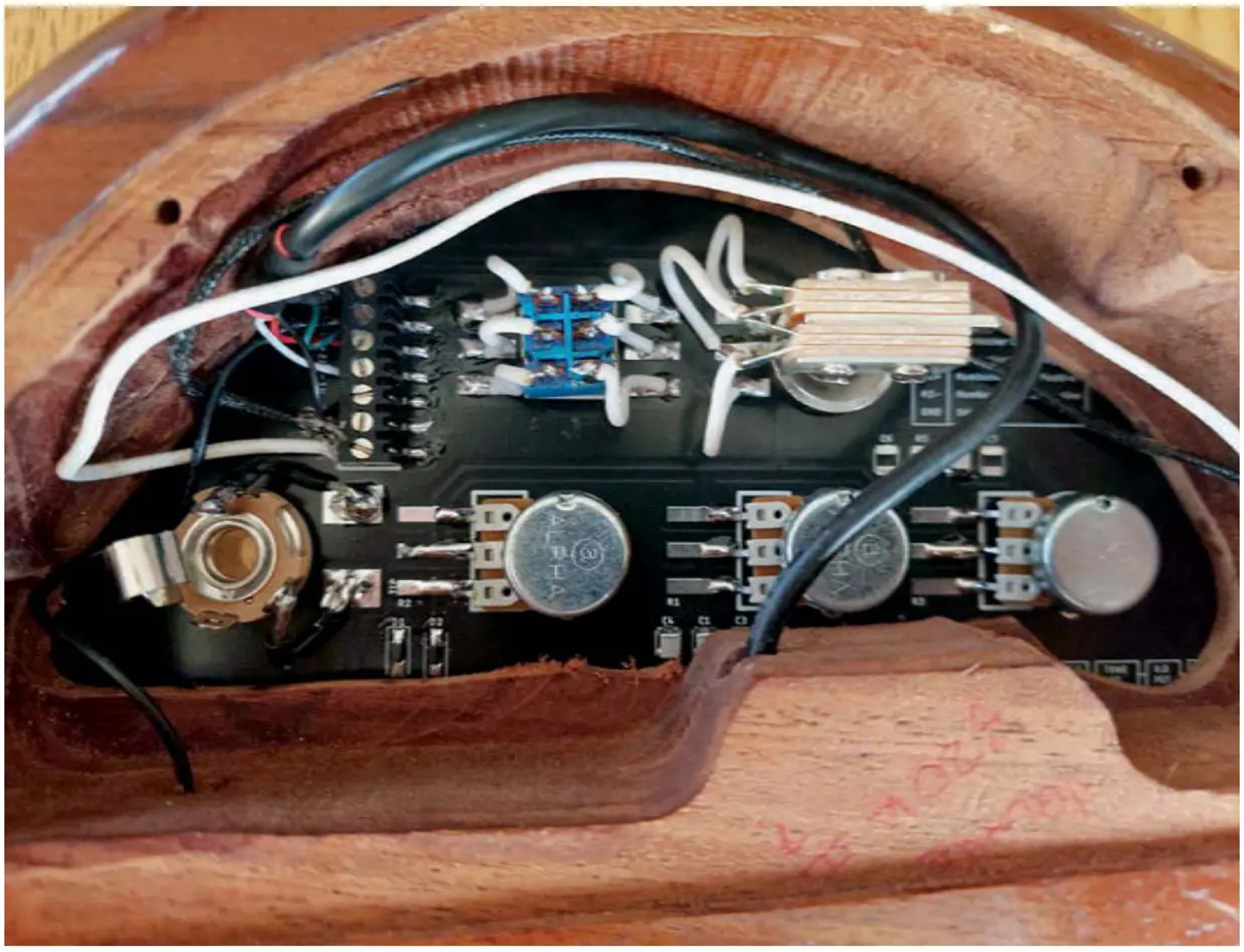


Photo 3: there's a fair bit of room inside the T-Style bass guitar body for the components on the underside of the PCB.

The signal is then fed to the traditional tone control, made from potentiometer VR1 (connected as a rheostat) and the four paralleled capacitors. This configuration forms a low-pass filter but in a slightly unusual way.

Because the pot is in series with the capacitor(s), it effectively works as a magnitude control. With the pot all the way up, the signal sees a high impedance and very little of the high end is shunted.

With the tone all the way down, the capacitor(s) are connected directly in parallel with the output, giving a strong high-frequency roll-off. The RC combination of the pickup source impedance and filter capacitance will loosely set the cutoff frequency. Still, with no buffering, there is a strong interaction with the other controls.

Because we prefer to use plastic film dielectric capacitors for linearity, the range of values available in the size used (SMD M3216/1206) is only up to about 1μF. The four footprints therefore allow for a broader range of capacitances and for tuning the sound by connecting smaller capacitors in parallel with larger ones.

A typical value is 47nF, but I prefer higher values to scoop out a bit more of the mid-range, so I use 200-220nF. In the prototype I built, I used two 100nF film capacitors in parallel, but I've specified a single 220nF in the parts list for simplicity.

The next control in the signal path is potentiometer VR2, which is a new feature: a passive overdrive/distortion control. It works similarly to the tone control, except there are back-to-back (inverse parallel) schottky diodes (D1, D2) in series with the rheostat.

When the pot is fully down, there

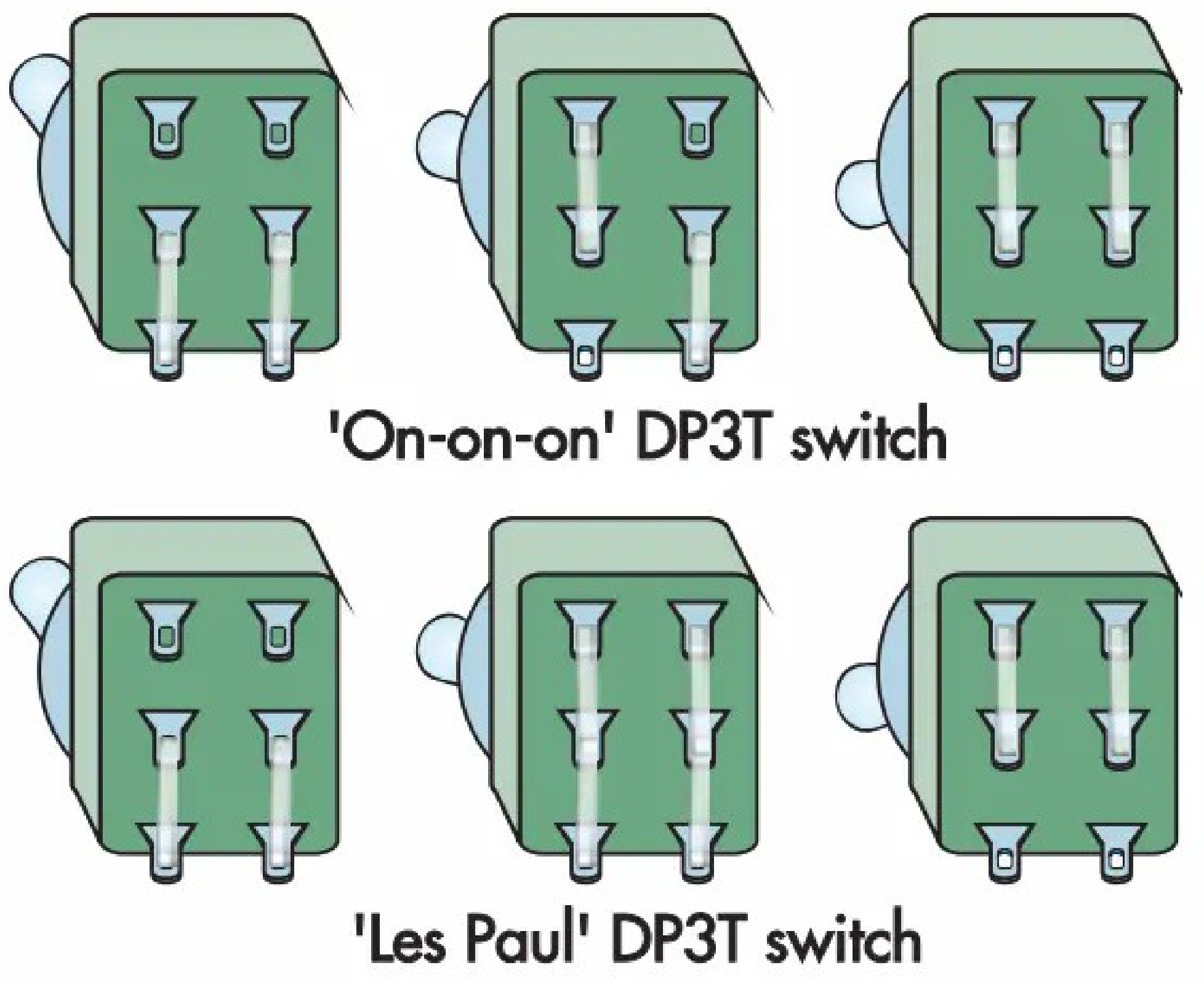
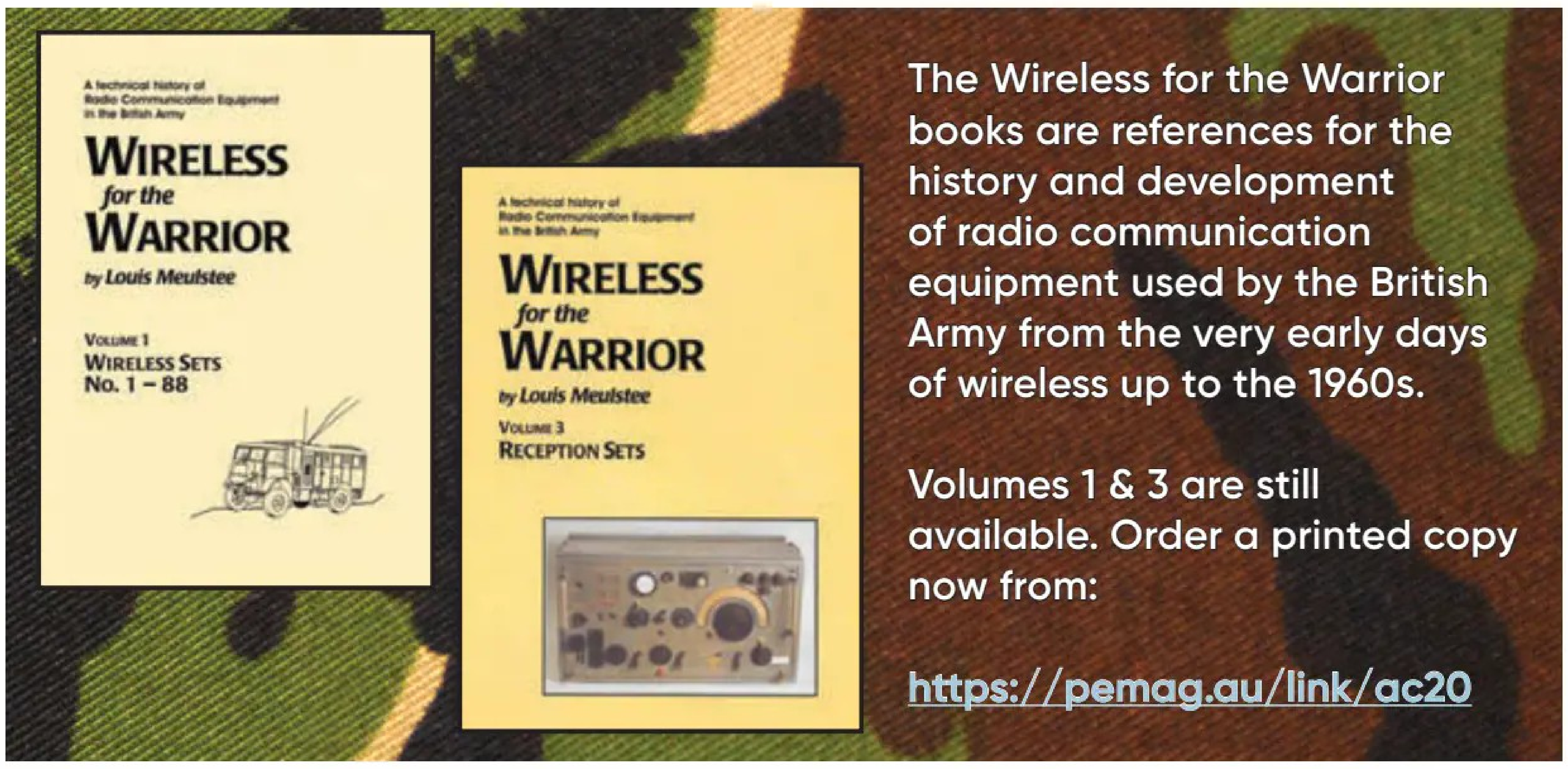


Fig.5: the unusual switching patterns of the "on/on/on" and "Les Paul" DP3T switches used in these circuits allows the coils to be used together or individually.


The Wireless for the Warrior books are references for the history and development of radio communication equipment used by the British Army from the very early days of wireless up to the 1960s.

Volumes 1 & 3 are still available. Order a printed copy now from:

https://pemag.au/link/ac20

is a high resistance in series with the diodes, so they have little effect on the signal.

When the pot is all the way up, the diodes are connected directly to the signal line. The voltage from the pickups is too low to fully forward bias these diodes, but even operating in their square law region, they introduce some nonlinearity to produce a subtle 'overdrive', a popular effect amongst guitarists.

The seasoned musician will note that the overdrive effect is applied before the volume control, so its impact won't be reduced if the volume is adjusted. This offers a useful contrast to other distortion sources, such as downstream foot pedals or amplifiers, which are mellowed by their input voltage level.

By combining the onboard distortion with downstream effects, the musician has the flexibility to dial up or down distortion from a mixture of sources.

Constructors might like to experiment with different combinations of diodes here; for instance, a single schottky diode would give asymmetric distortion, while back-to-back combinations of small signal diodes like the 1N4148WS would provide a more mellow effect. You could even have a combination, with one 1N4148WS and one schottky diode facing in either direction.

The final potentiometer, VR3, is a traditional volume control with a twist. In a conventional circuit, the output signal is simply tapped off the pot's wiper. But this arrangement has a drawback: as the volume is turned down, the pot resistance appears in series with the output. When connected to an amplifier via a coaxial cable, this resistance forms an unwanted low-pass filter (with the cable parasitic capacitance), reducing upper frequencies.

In this updated circuit, a 1nF capacitor is placed in parallel with the volume control to 'bleed' additional treble into the output as the volume is turned down, compensating for the undesirable tone loss.

Additional series and parallel footprints R5, C6 and C7 are provided for other combinations of capacitance or resistance to do this job. For instance, 100kΩ || 1nF may sound more linear as the control is turned down.

The output signal appears at the output jack, CON, a ¼-inch (6.35mm) TS socket to suit a standard instrument cable.

The PCB is secured to the front of the instrument using 3mm stainless steel self-tappers. They mount through 3mm plated through-holes and secure directly into the timber. I prefer plated holes for mechanical mounting, as they are a bit more hardy than bare fibreglass against the metal screw threads.

A standard 1.6mm thickness PCB with black solder mask is best; any thinner would be too flimsy, any thicker would present too much of a lip. We will be supplying boards with a lead-free HASL finish (basically tin plating) as the solder will be on the outside of the guitar. If a gold finish would suit your guitar, you could go for an ENIG finish, although it will make the board considerably more expensive.

The Fender Jazz Bass

Many of the features of this T-style bass circuit can be applied to more

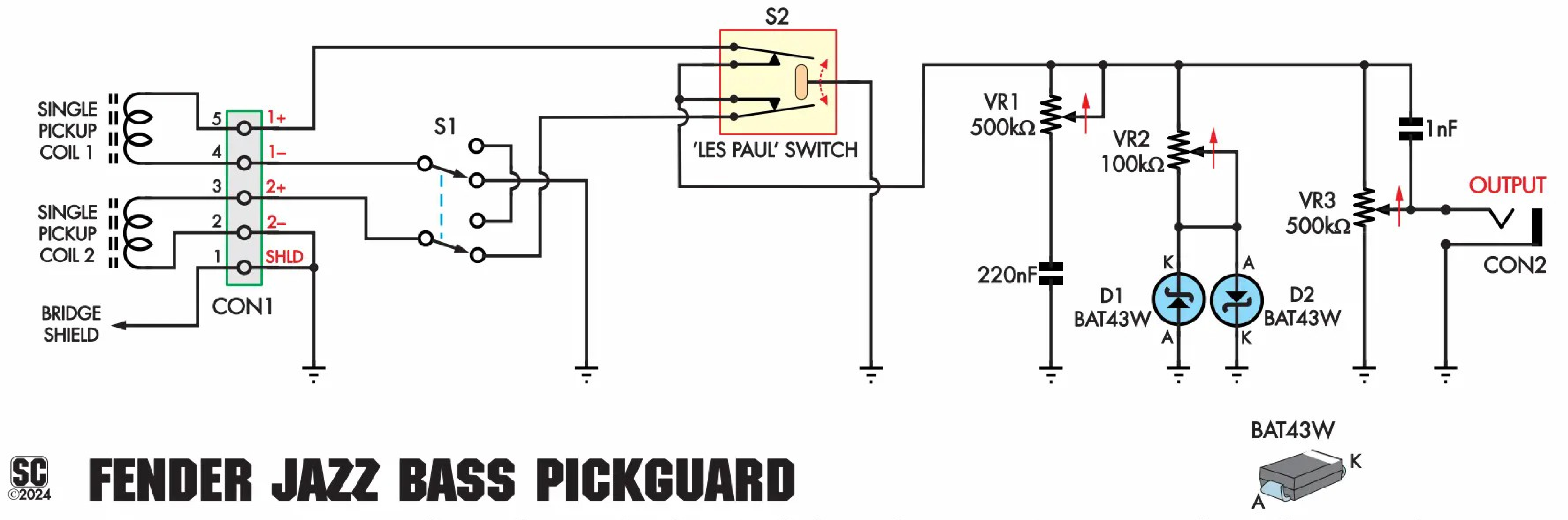


Fig.6: my Jazz Bass circuit is similar to the one for the T-Style bass shown in Fig.4, except the pickup switching is simpler because both pickups are single-coil types.

Fender
JAZZ BASS

Photo 4: a Fender Jazz electric bass. ***Source:*** ***www.megamusiconline.com.au/product/fender-american-performer-jazz-bass-guitar-rosewood-fretboard-3-colour-sunburst/***

common instruments. One of the most popular bass guitars is the Fender Jazz Bass (Photo 4), which has two single-coil pickups. Jazz basses have been played extensively by legends like Jaco Pastorius, John Paul Jones of Led Zeppelin, Flea of Red Hot Chili Peppers, Adam Clayton of U2 and Geddy Lee of Rush.

There are also clones of the Jazz Bass (and the other guitars listed below) that would likely fit my new pickguard designs, possibly with slight modifications to the inside of the body.

Traditionally, the two pickups would be wired to individual volume control potentiometers with a shared common tone control. My new arrangement is shown in Fig.6.

The negative of the second coil and the bridge shield are connected directly to circuit ground, similarly to the T-Style circuit from above. The first coil's negative and the second coil's positive are routed to S1, a DPDT toggle switch that provides series/individual switching for the two coils. With S1 in the position shown, S2 allows the player to select either coil or both in parallel.

Parallel is the standard configuration for a Jazz bass, while series is a new mode that will give a stronger and fuller tone. Series switching with two single-coil pickups is a rare configuration but, in my opinion, heavily underrated. On a genuine Fender, it can give the player a beefier tone that is more akin to a humbucker. It can compensate for low-cost pickups, which tend to sound thin on a cheap imitation.

Photo 5 shows how this special 'Les Paul style' switch works; with the toggle in the central position, all contacts are closed.

Moving it to one side opens the contacts on the opposite side, while leaving the set on the same side closed. In its standard configuration, that lets you choose one pickup, the other or both in parallel.

With S1 in the standard parallel mode, S2 can select between the neck pickup only, bridge pickup only, or both pickups in parallel. With S1 in the series position, S2 selects between both in series or mute.

Mute can be helpful for live work, to avoid unwanted sounds when moving around on stage between songs, or it can be rapidly switched on and off to give a tremolo-style effect.

The signal is then sent to the tone control potentiometer, VR1. A single capacitor is used here (220nF recommended), as there is no space for more footprints. Following this is pot VR2, the passive overdrive with dual schottky diodes.

After that is pot VR3, the volume control with a single 1nF treble

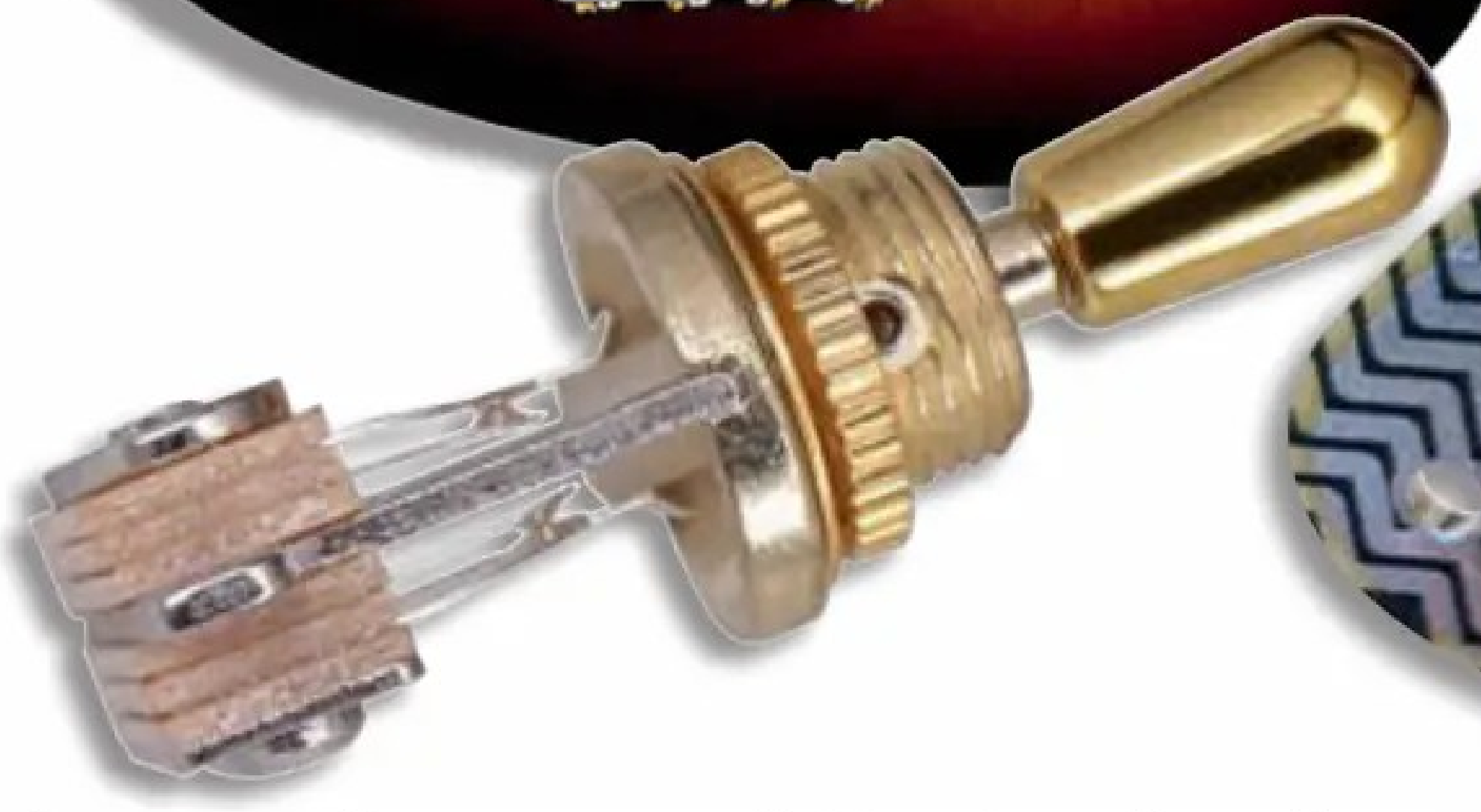
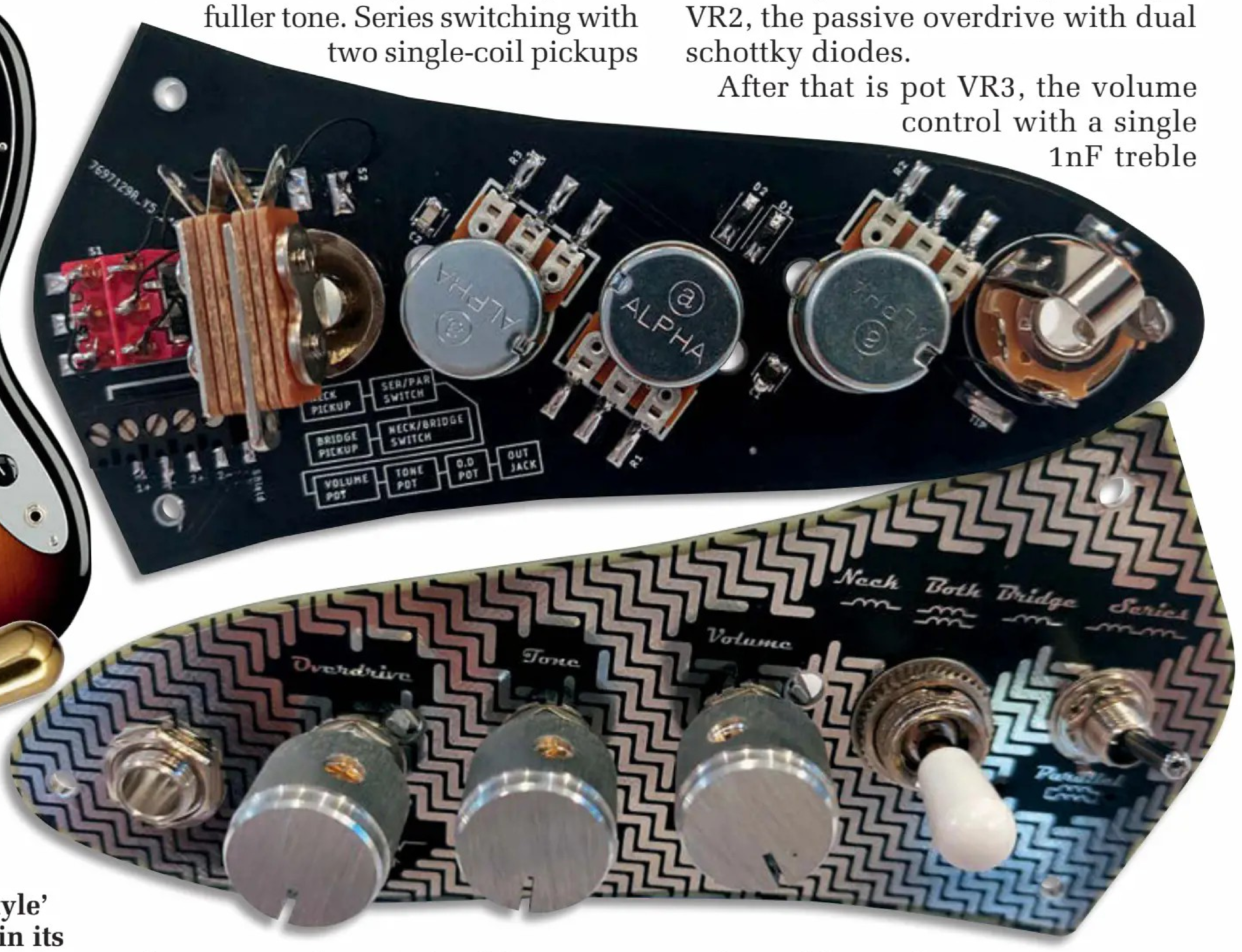


Photo 5: the unusual 'Les Paul style' DPDT switch closes all contacts in its central position, rather than opening them all, as in a normal DPDT centre-off switch.



Photos 6 & 7: the outside of the Jazz Bass pickguard has a pleasing zigzag pattern in tin along with the necessary labels. The other side of the pickguard is where all the components are mounted.

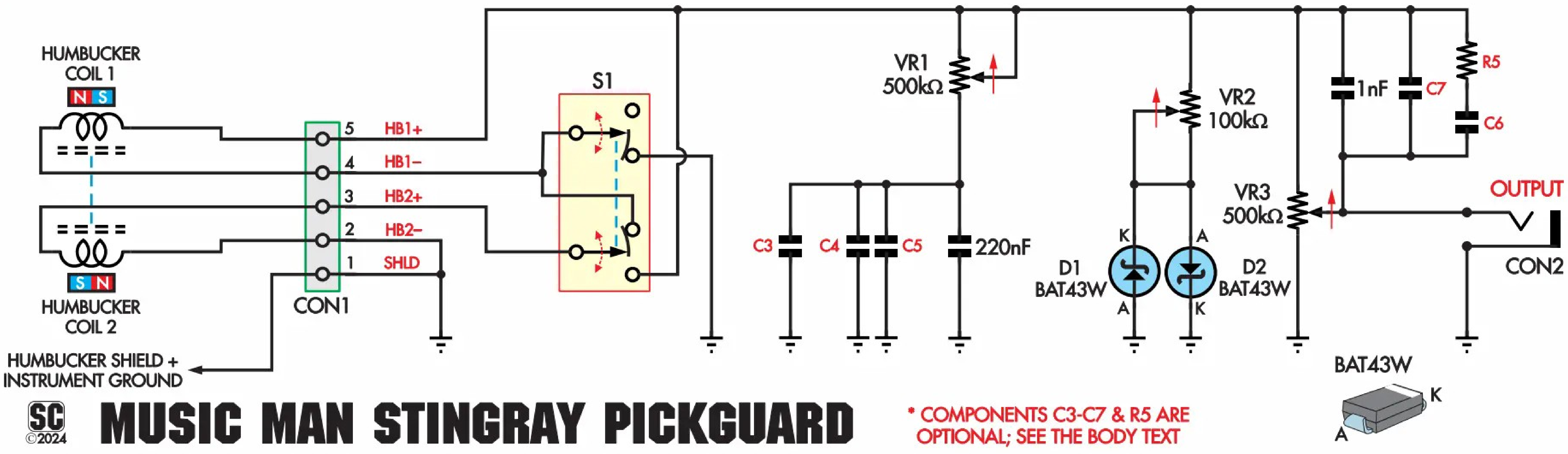


Fig.7: the Music Man Stingray circuit is virtually identical to the one for the J&D Luthiers T-Style bass, except that there is no switch S2 as it doesn't have a neck pickup.

bleed capacitor. Again, there is no space for additional footprints or series/parallel combinations as there was on the T-Style circuit.

The entire PCB fits in place of the existing chrome-plated controls (visible in Photo 4) and is screwed down to the body. If your instrument doesn't quite fit the new controls (shown in Photos 6 & 7), additional space can be made by filing, rasping, or routing out a larger body cavity.

Music Man Stingray

Another very popular bass guitar, the Stingray (Photo 8), was designed by Leo Fender after he sold his interest in his founding business, the Fender music instrument company.

Unusually for a Fender design, it features a humbucker pickup. Legends who have used this bass include Cliff Williams of AC/DC and John Deacon of Queen on tracks like Another One Bites the Dust.

The standard Stingray was the first electric instrument to feature active electronics, with a volume control and a two-band EQ powered by a 9V battery. Some models add a knob to provide a three-band EQ. Cheaper copies of the Stingray will come with a similar circuit to the Jazz Bass: two volume control potentiometers for each of the humbucker coils and a common tone control.

My replacement circuit is shown in Fig.7 and Photos 9 & 10. The bridge shield and humbucker coil two negative are connected to circuit ground. Like the humbucker wiring in the T-Style circuit, humbucker coil two positive and humbucker coil one negative connect to S1, a DP3T toggle switch with the special on/on/on switching pattern.

This provides the same series/split/parallel switching for the humbucker as the T-Style circuit, with the same tonal flexibility.

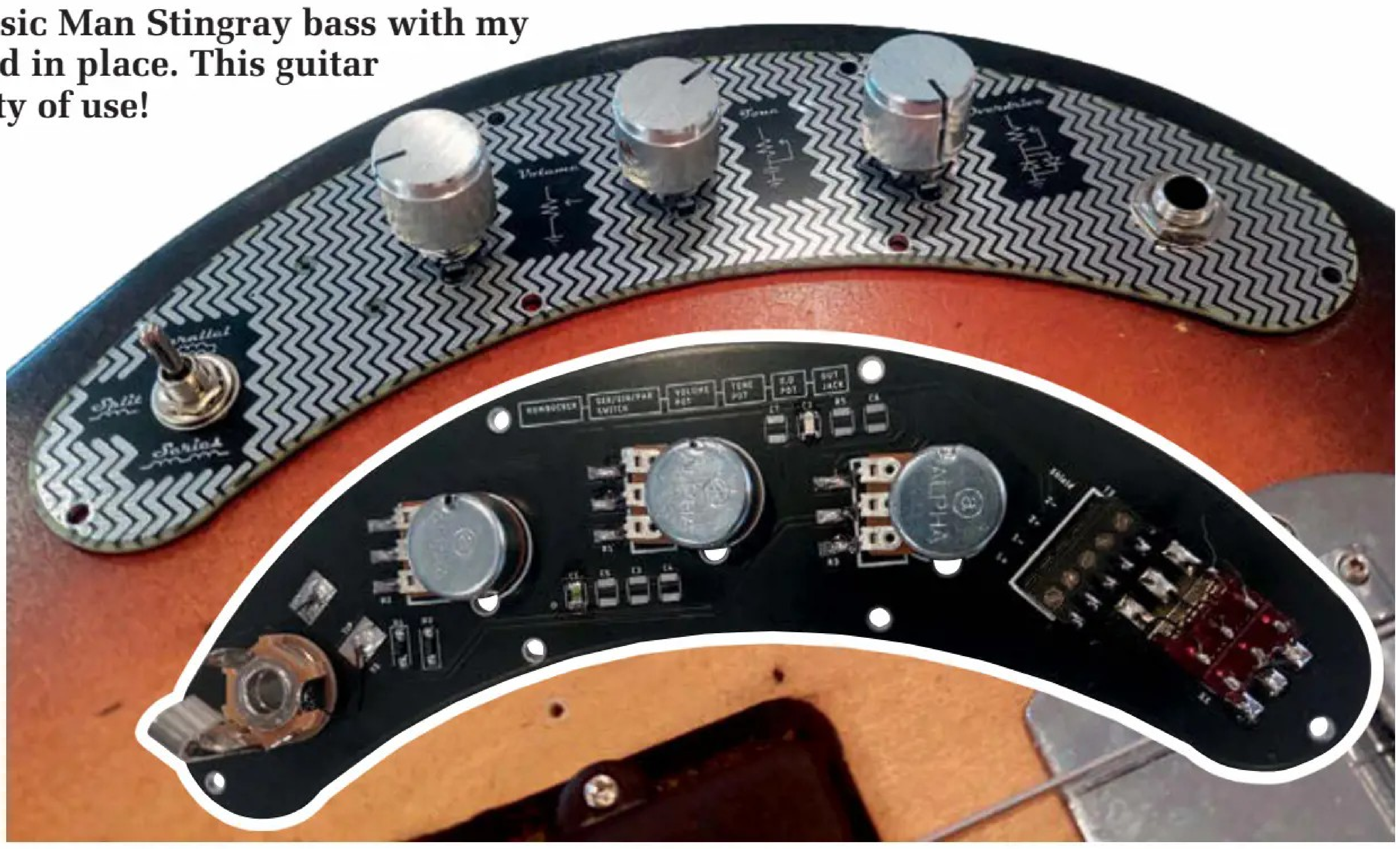
When in split mode, only a single coil is active. Depending on which coil the user prefers, this can be the one closest to the bridge or nearest to the neck. It can be changed by simply swapping the wiring of coil 1 with coil 2 in the screw terminal.

As the two pickups are mounted so close to each other, there is only a very minor difference in sound between the two, but the bridge coil will be marginally 'brighter', so I prefer to use it.

The signal from S1 connects with the humbucker coil one positive and is sent to the tone control formed by potentiometer VR1 and up to four paralleled capacitors.

In this example, a single 220nF film

◀ **Photo 8: a Music Man Stingray bass with my new pickguard in place. This guitar has seen plenty of use!**



Photos 9 & 10: a close-up of the controls on the Stingray, showing how the orientation of the labels makes sense for the guitar player. Generally, only capacitors C1 and C2 are required, as shown here, but the extra pads give you more options.

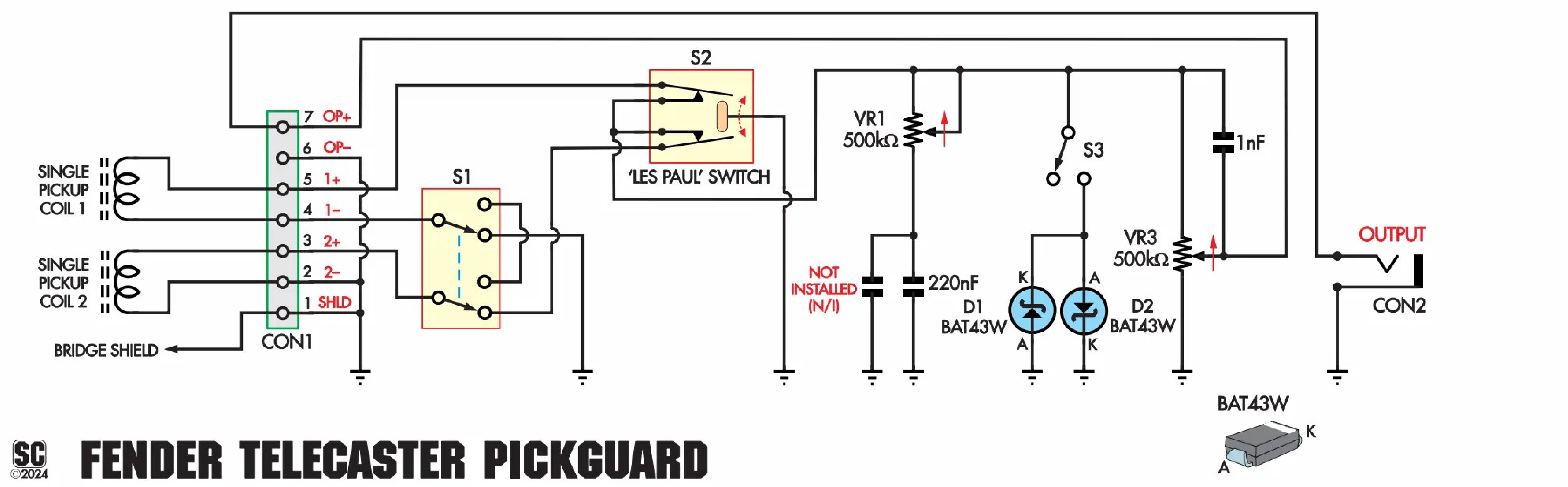

SINGLE PICKUP COIL 1
SINGLE PICKUP COIL 2
BRIDGE SHIELD
CON1
7 OP+
6 OP–
5 1+
4 1–
3 2+
2 2–
1 SHLD
S1
S2
'LES PAUL' SWITCH
VR1 500kΩ
NOT INSTALLED (N/I)
220nF
S3
K
A
D1 BAT43W
D2 BAT43W
VR3 500kΩ
1nF
OUTPUT
CON2
BAT43W
SC ©2024
FENDER TELECASTER PICKGUARD

Fig.8: while the other three circuits were for bass guitars, this one is for a Fender Telecaster standard electric guitar. It's the simplest of the four due to the minimal space available on the guitar, with pickup switching, tone and volume controls and a simple on/off overdrive option.

cap is loaded. After this is the overdrive control, formed by potentiometer VR2 and inverse-parallel schottky diodes D1& D2.

The final control is the volume control, VR3, with the treble bleed circuitry already described. A single 1nF capacitor (C2) is used in my prototype, but footprints R5, C6, and C7 are also provided to give flexibility to the builder.

Fender Telecaster

The previous circuits have all been designed for electric bass guitars but can also be applied to electric guitars. The Fender Telecaster (Photo 11) is one of the oldest and most popular electric guitars, and its two-pickup combination can work with a circuit similar to that of the Jazz Bass. There are too many famous Telecaster players to list, including Jimmy Page, Keith Richards, Bruce Springsteen, George Harrison and Muddy Waters.

The standard controls on a telecaster are a three-way pickup selector switch (neck/bridge/both) and the typical volume and tone controls. These can be replaced with the circuit shown in Fig.8, Photo 12 & Photo 13.

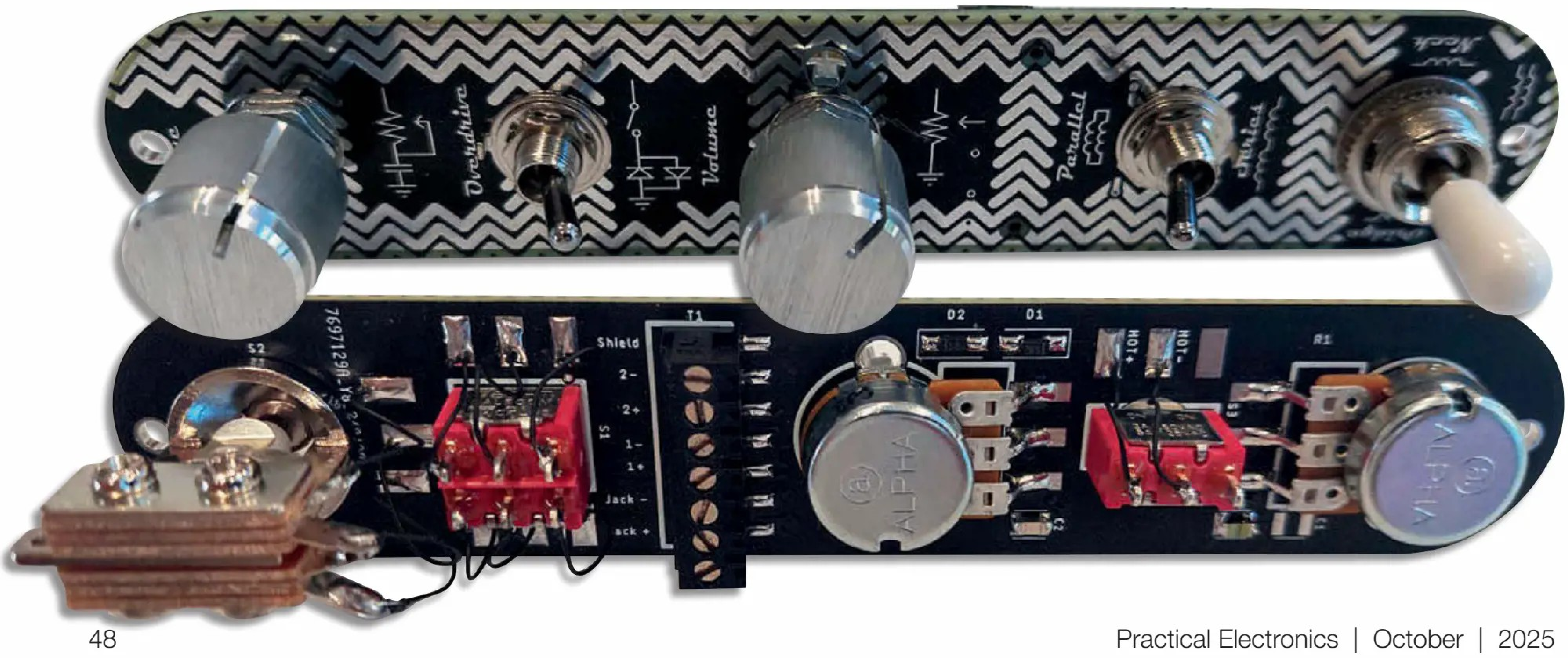
The output jack negative, the second single coil negative and the bridge shield connect directly to circuit ground. The negative of the first single coil and the positive of the second single coil are routed to S1, a DPDT toggle switch providing series/parallel switching for the two coils.

The output from S1 is sent to S2, a three-way bridge/neck selector switch. With S1 in the standard parallel mode, S2 can select between the neck pickup only, bridge pickup only, or both pickups. With S1 in the series position, S2 selects between on and mute.

The signal is then sent to the tone control potentiometer, VR1. Two footprints are provided for capacitors. After this is the passive overdrive, which differs from other circuits in its use of a switch rather than a potentiometer.

Due to space constraints, the dual schottky diodes are simply switched in or out of circuit. An SPST or SPDT toggle switch can be used here. After that is pot VR3, the volume control, with a single 1nF treble bleed capacitor.

Photos 12 & 13: once your Telecaster pickguard has been assembled and the wires added, it should look like this, ready to install in the guitar. The pickguard is packed with controls and has a zigzag pattern to add a bit of interest.


Overdrive
Volume
Parallel
Series
Neck
Bridge
S2
S1
T1
Shield
2-
2+
1-
1+
Jack -
Jack +
D2
D1
HOT+
HOT-
R1
ALPHA

Electric Guitar/Bass Pickguards

Photo 11: a Fender Telecaster electric guitar. ***Source:*** *www.keyboardcorner.com.au/fender-player-telecaster-pau-ferro-fingerboard-3-color-sunburst/*

This circuit does not include a ¼-inch output jack, as the Telecaster mounts this off-board on the bottom edge of the body. The output signal instead connects back to the 7-way screw terminal, where it can connect to flying leads that wire down to the output jack.

Construction

All versions can be approached similarly, referring to the photos presented so far and the relevant PCB overlay diagram (one of Figs.9-12). Begin by mounting the toggle switch(es) and securing them to the PCB with the supplied washer and nut.

I prefer a flush mount; if the toggle stands too proud for your liking, an additional nut can be placed on the back of the switch to adjust the length of shaft that protrudes through the PCB.

Tin the SMD pads and toggle switch terminals with solder, then run short lengths of fine-gauge wire from the pads to the terminals. The DPDT and DP3T switches require all six connections, while the SPST/SPDT on the Telecaster circuit only needs the two connections as marked.

Continue by mounting the ¼-inch jack (excluding the Telecaster). Secure it to the circuit board using the supplied washer and nut, like the toggle switches. Tin the two terminals and the pads and make connections with the fine gauge wire. The hot pad connects to the tip of the instrument cable, the longest conductor on the jack. The ground

Parts List – Electric & Bass Guitar Pickguards

Parts common to all versions

1 2 500kΩ logarithmic taper single-gang 16mm spline shaft potentiometers (VR1, VR3)
1 6.35mm switched SPST mono jack socket (CON2) *
3 16mm aluminium ¼-inch shaft grub screw knobs
2 BAT43W 30V 200mA SMD schottky diodes, SOD-123 (D1, D2) **[DigiKey, Mouser etc]**
1 220nF 16V metallised plastic film or NP0/C0G ceramic capacitor, M3216/1206 size **[DigiKey ECP-U1C224MA5]**
1 1nF 100V metallised plastic film or NP0/C0G ceramic capacitor, M3216/1206 size **[DigiKey ECW-U1102JX5]**
1 1m length of black light-duty hookup wire
** omit jack socket and one knob for Telecaster*

Fender Jazz Bass specific parts

1 double-sided PCB with black solder mask coded 23109241, 127 × 105.5mm
1 5-way SMD screw terminal, 3.5mm pitch (CON1) **[DigiKey 2383942-5]**
1 DPDT solder tail mini toggle switch (S1)
1 "Les Paul 3 Way Selector" DP3T switch (S2) **[AliExpress 1005001900886767]**
1 100kΩ linear taper single-gang 16mm spline shaft potentiometer (VR2)

J&D Luthiers T-Style specific parts

1 double-sided PCB with black solder mask coded 23109242, 200.5 × 87.5mm
1 7-way SMD screw terminal, 3.5mm pitch (CON1) **[DigiKey 2383942-7]**
1 'on-on-on' DPDT miniature toggle switch with solder tags (S1) **[AliExpress 1005008032212175]**
1 "Les Paul 3 Way Selector" DP3T switch (S2) **[AliExpress 1005001900886767]**
1 100kΩ linear taper single-gang 16mm spline shaft potentiometer (VR2)

Music Man Stingray specific parts

1 double-sided PCB with black solder mask coded 23109243, 190 × 71mm
1 5-way SMD screw terminal, 3.5mm pitch (CON1) **[DigiKey 2383942-5]**
1 'on-on-on' DPDT miniature toggle switch with solder tags (S1) **[AliExpress 1005008032212175]**
1 100kΩ linear taper single-gang 16mm spline shaft potentiometer (VR2)

Fender Telecaster specific parts

1 double-sided PCB with black solder mask coded 23109244, 28.5 × 148mm
1 7-way SMD screw terminal, 3.5mm pitch (CON1) **[DigiKey 2383942-7]**
1 DPDT solder tail mini toggle switch (S1)
1 "Les Paul 3 Way Selector" DP3T switch (S2) **[AliExpress 1005001900886767]**
1 SPDT solder tail mini toggle switch (S3)

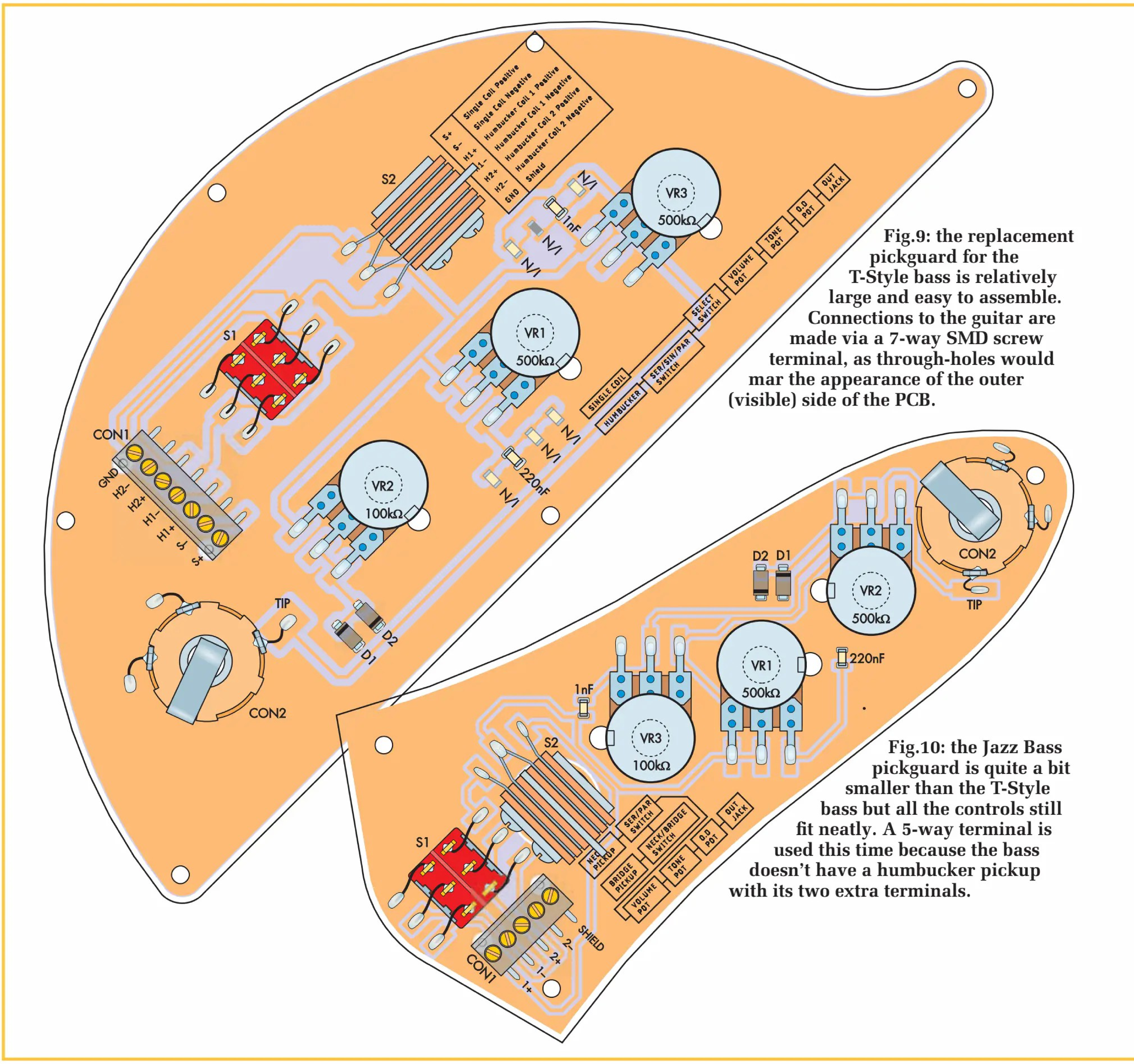

S+ Single Coil Positive
S– Single Coil Negative
H1+ Humbucker Coil 1 Positive
H1– Humbucker Coil 1 Negative
H2+ Humbucker Coil 2 Positive
H2– Humbucker Coil 2 Negative
GND Shield
S2
S1
CON1
GND H2– H2+ H1– H1+ S– S+
VR3 500kΩ
VR1 500kΩ
VR2 100kΩ
1nF
220nF
N/I
D2 D1
TIP
CON2
OUT JACK
O.D POT
TONE POT
VOLUME POT
SELECT SWITCH
SER/SIN/PAR SWITCH
SINGLE COIL
HUMBUCKER
CON2
TIP
VR2 500kΩ
D2 D1
VR1 500kΩ
220nF
1nF
VR3 100kΩ
S2
S1
CON1
1+ 1– 2+ 2– SHIELD
NECK PICKUP
BRIDGE PICKUP
SER/PAR SWITCH
NECK/BRIDGE SWITCH
VOLUME POT
TONE POT
O.D POT
OUT JACK

Fig.9: the replacement pickguard for the T-Style bass is relatively large and easy to assemble. Connections to the guitar are made via a 7-way SMD screw terminal, as through-holes would mar the appearance of the outer (visible) side of the PCB.

Fig.10: the Jazz Bass pickguard is quite a bit smaller than the T-Style bass but all the controls still fit neatly. A 5-way terminal is used this time because the bass doesn't have a humbucker pickup with its two extra terminals.

connection is to the jack's sleeve.

Now mount the potentiometers. The Volume and Tone controls are nominally 500kΩ audio taper, but other values can be substituted if the builder prefers.

For instance, 250kΩ and 1MΩ pots are also commonly used on many instruments. Generally speaking, higher values will give a brighter voicing to the instrument but will also increase the circuit's output impedance.

The potentiometers are mounted flush against the PCB, with the provided keyway holding the control in place to prevent unwanted body rotation.

The terminals can be bent down towards the pads on the PCB and soldered with a blob. Begin by applying solder to the legs of the pot and wait for the solder to reflow down onto the pads for a good connection.

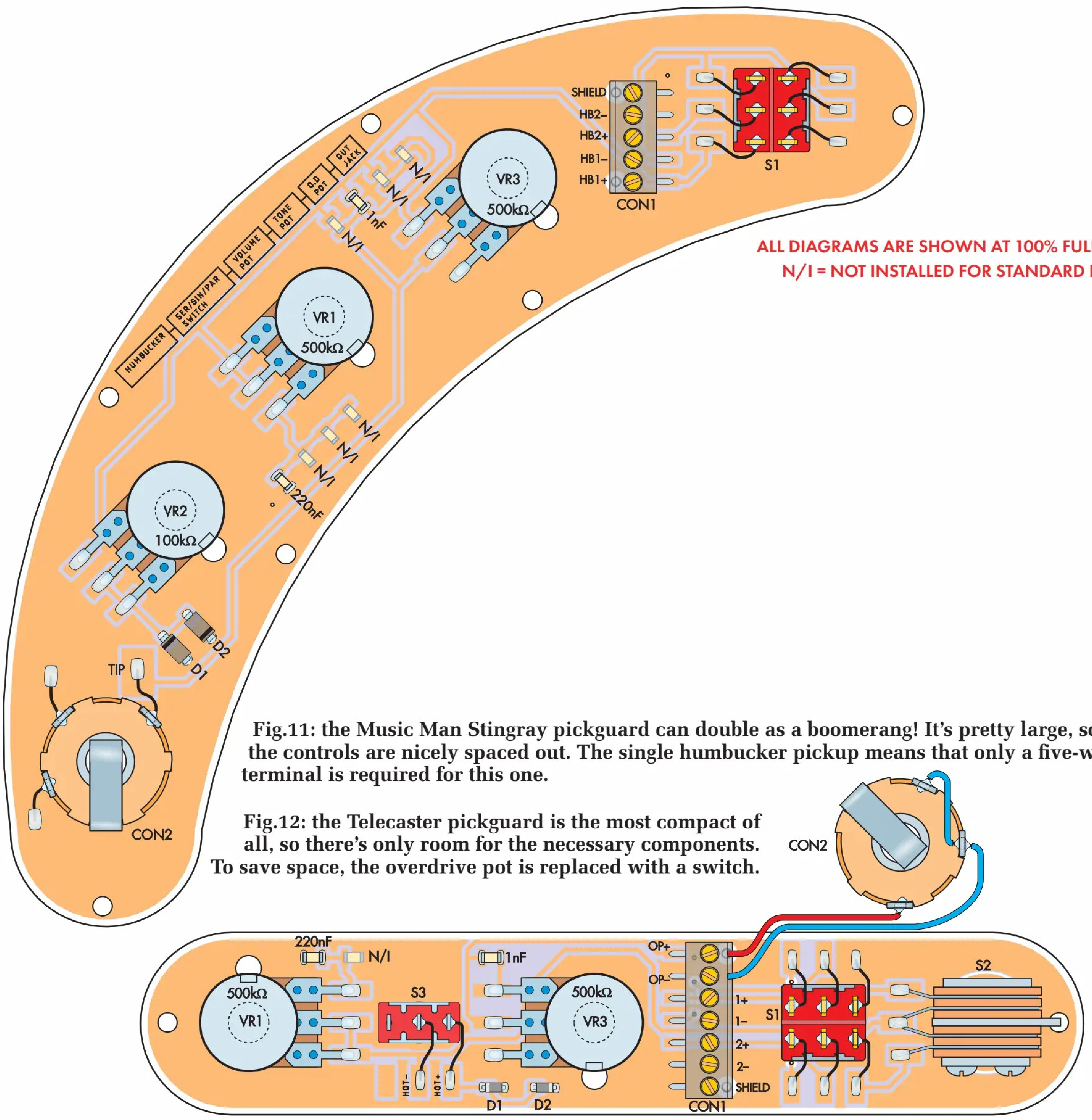
Finish the potentiometers by mounting the brushed aluminium knobs to the shaft. If the pot shaft is too long, it can be cut with a hacksaw and filing to give a flush mount. The knobs secure to the shaft by tightening the grub screw.

Next, mount the SMD screw terminal. Solder each leg individually in a similar way to the pot legs; flow solder onto the legs and wait for it to reflow onto the pads of the PCB.

Finally, the passives can be soldered to the pads on the circuit board. I used plastic film caps throughout, as plastic dielectrics are highly linear. Finish by soldering the schottky diodes with opposite orientations.

Editor's note: C0G ceramics (also called NP0) are more linear than many common film caps, such as polyester types, so they might give a more neutral sound.

The pickups can now be wired to the screw terminal, and the entire assembly mounted to the body of the instrument with 3mm wood screws. Happy playing! PE



ALL DIAGRAMS ARE SHOWN AT 100% FULL SIZE
N/I = NOT INSTALLED FOR STANDARD BUILD

Fig.11: the Music Man Stingray pickguard can double as a boomerang! It's pretty large, so all the controls are nicely spaced out. The single humbucker pickup means that only a five-way terminal is required for this one.

Fig.12: the Telecaster pickguard is the most compact of all, so there's only room for the necessary components. To save space, the overdrive pot is replaced with a switch.

The Fox Report

... continued from page 41.

Flyby11 (now called Flyoobe) is a free software tool that helps owners of Window 10 PCs bypass the upgrade barriers involving TPM, Secure Boot and CPU type imposed by Microsoft.

It is not a push-button-and-go installation tool; it's an assistant that removes the restrictions during the upgrade. Flyoobe is not for novices. It may or may not wipe previously installed programs and settings, but with the option to keep previously stored data. You need your wits about you before pressing Continue.

I used Flyoobe with an old, underpowered, treacle-slow laptop that had previously refused all forced upgrade tricks. In my case, Flyby/Flyoobe worked by using the installation mode created by Microsoft to install the Windows Server variant of Windows.

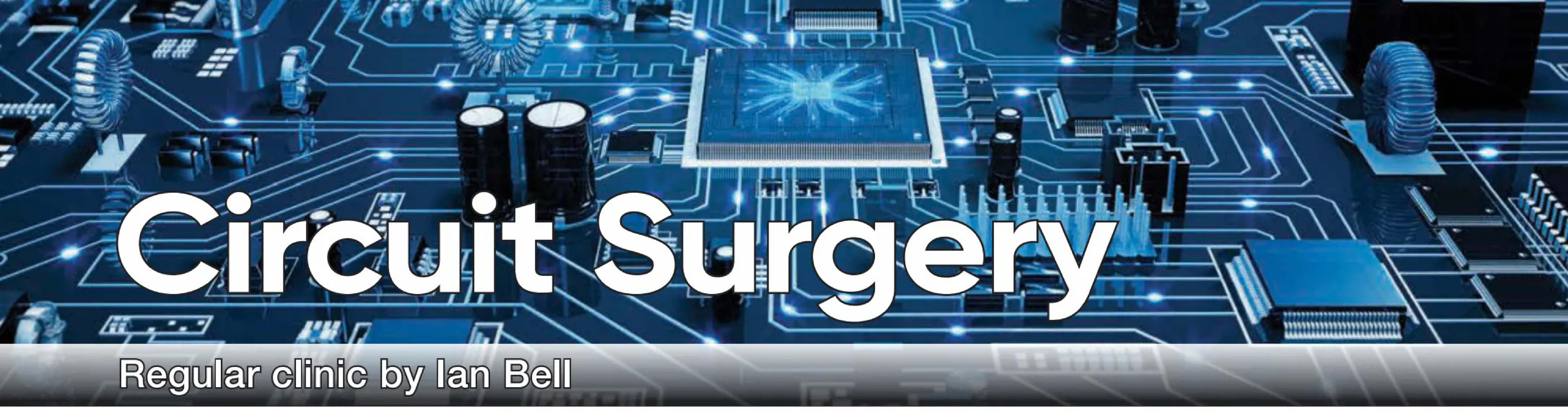
Although the tool says it is installing Windows Server, it actually downloads the normal Windows 11 ISO, burns it to an 8GB USB stick and installs bog-standard Windows.

The process is lengthy, so I left the last stages running overnight. In the morning, the laptop astonished me by starting up with a Windows 11 Home Version 24H2 nicely installed. My data remained intact, but the desktop program and app icons no longer worked.

I needed to re-connect to Wi-Fi and re-sign in to Google and Gmail. The installed Windows 11 then needed a shedload of Windows 11 updates. Almost all programs and apps needed reinstalling but that's a small price to pay.

What matters is that Flyoobe works. It has achieved what Microsoft is trying to prevent, and can save lower-end PCs from the scrapheap. If anything, my 'new' Windows 11 laptop runs faster than my 'old' Windows 10 machine.

Congratulations to the developer – who identifies only as Belim and asks for (well-deserved) donations via PayPal. You can get it from https://github.com/builtbybel PE



Circuit Surgery

Regular clinic by Ian Bell

Measuring the frequency response of a circuit or device using with a PC sound card, part 1

Last month, we reached the end of our series on digital signal processing (DSP), which has covered the various parts of a generic DSP system and the basics of their theory of operation. In the most recent few articles, we described the implementation of a basic digital filter using a microcontroller.

In the last issue, we showed oscilloscope traces for a couple of input and output signals to confirm a low-pass filter example did attenuate higher frequencies. However, what we really want to do to evaluate a filter is to plot its measured frequency response.

This month, we will start to look at frequency response measurement in general, not just for the DSP example. The frequency response of a circuit is the variation of its behaviour with frequency. If the input is X and the output is Y, we could write $Y(f) = H(f)X(f)$, where H is the transfer function – the relationship between the input and output signals.

We write $Y(f)$, $H(f)$ and $X(f)$ to show that the signals (which could be voltage or current) and the transfer function vary with frequency (f). The term "frequency response" often refers to graphs showing how the transfer function $H(f)$ varies with frequency.

Frequency response graphs often require a lot of data points, so it can be time-consuming and tedious to measure and calculate each point manually. Professional labs use test gear that can automate the measurement; many mid-to-high-end oscilloscopes now have this capability, but it is less common in older and lower-cost models.

Alternatively, a separate signal generator and oscilloscope can be coordinated by software to automate measurements (eg, using the SCPI protocol).

Not all readers will have even a basic oscilloscope, which could be used for manual measurements. It is possible to use a digital multimeter (DMM) to measure the gain response, but only if the meter can measure AC voltage over the required frequency range; basic models typically have limited capabilities in this respect. An audio millivoltmeter is a more appropriate instrument for this task.

You would also need a signal generator to create the sinewave input. Many digital oscilloscopes these days come with built-in arbitrary waveform generators.

If you do not have suitable test gear, there is a good alternative – use the "sound card" of a PC or laptop to generate the input signal and analyse the response. Fortunately, free software is available that provides this facility.

These programs are mainly aimed at acoustic measurements (room acoustics, loudspeakers etc), not directly testing electronic circuits, but they can be used for this purpose. We will look at how to do this.

By the way, we refer to the audio input/output circuitry as a "sound card" because it used to be the case that computers had no such circuitry, so you needed an add-on card (circuit board) to provide these functions. These days, almost all computers have analog audio input/output circuitry integrated onto their main boards, but the name has stuck.

Before getting into the specifics of this process, we shall look at the basics of frequency response measurement, both the theory and some general practical considerations.

Frequency response concepts

This expression, $Y(f) = H(f)X(f)$, implies linearity. That is, at a **given frequency**, the value of H is fixed – it varies with f, but not with other things such as time or any other characteristic of the input signal, such as its amplitude.

H not changing with X (ie, only with f) means that for a particular frequency (f_1, say) we could write $X=H_{f_1}Y$ and a graph of X (input) against Y (output) at that frequency would be a straight line (hence 'linear').

Linearity in combination with H not varying with time means we have what is called a linear time-invariant (LTI) system. A lot of theory behind electronics design is based on the assumption that circuits like amplifiers and filters are LTI systems.

Given that we are interested in characterising a circuit's behaviour with frequency, it makes sense to analyse or measure circuits by applying signals containing a single frequency. We can repeat this for as many frequencies as we need to build a picture of the frequency response. This is not the only way to obtain a frequency response, but it is the most straightforward to think about it.

Sinusoidal waveforms are the only waveforms that contain just a single frequency – other wave shapes have content at multiple frequencies. This is the basis of Fourier theory, which states that any waveform can be decomposed into purely sinusoidal components.

So, we can obtain a frequency response (in theory, simulation and practical

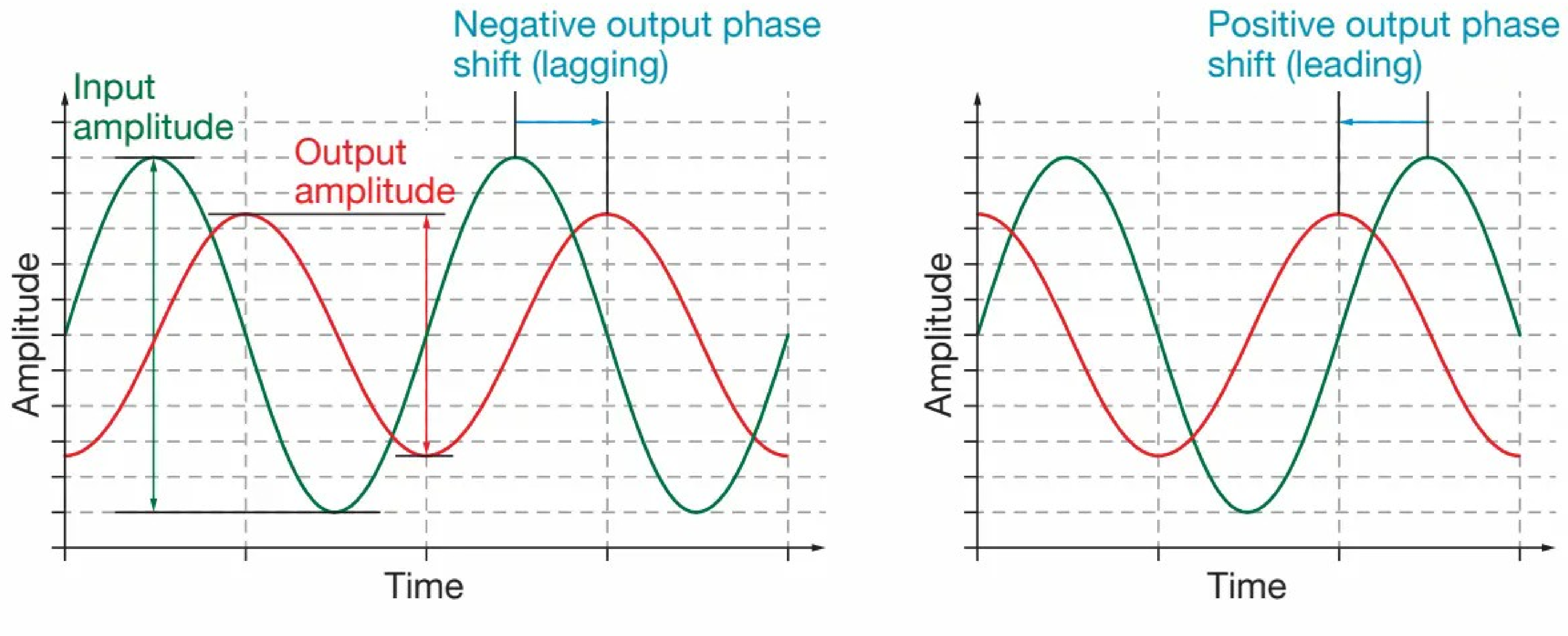


Fig.1: possible amplitude and phase shifts of some sample sinewaves.

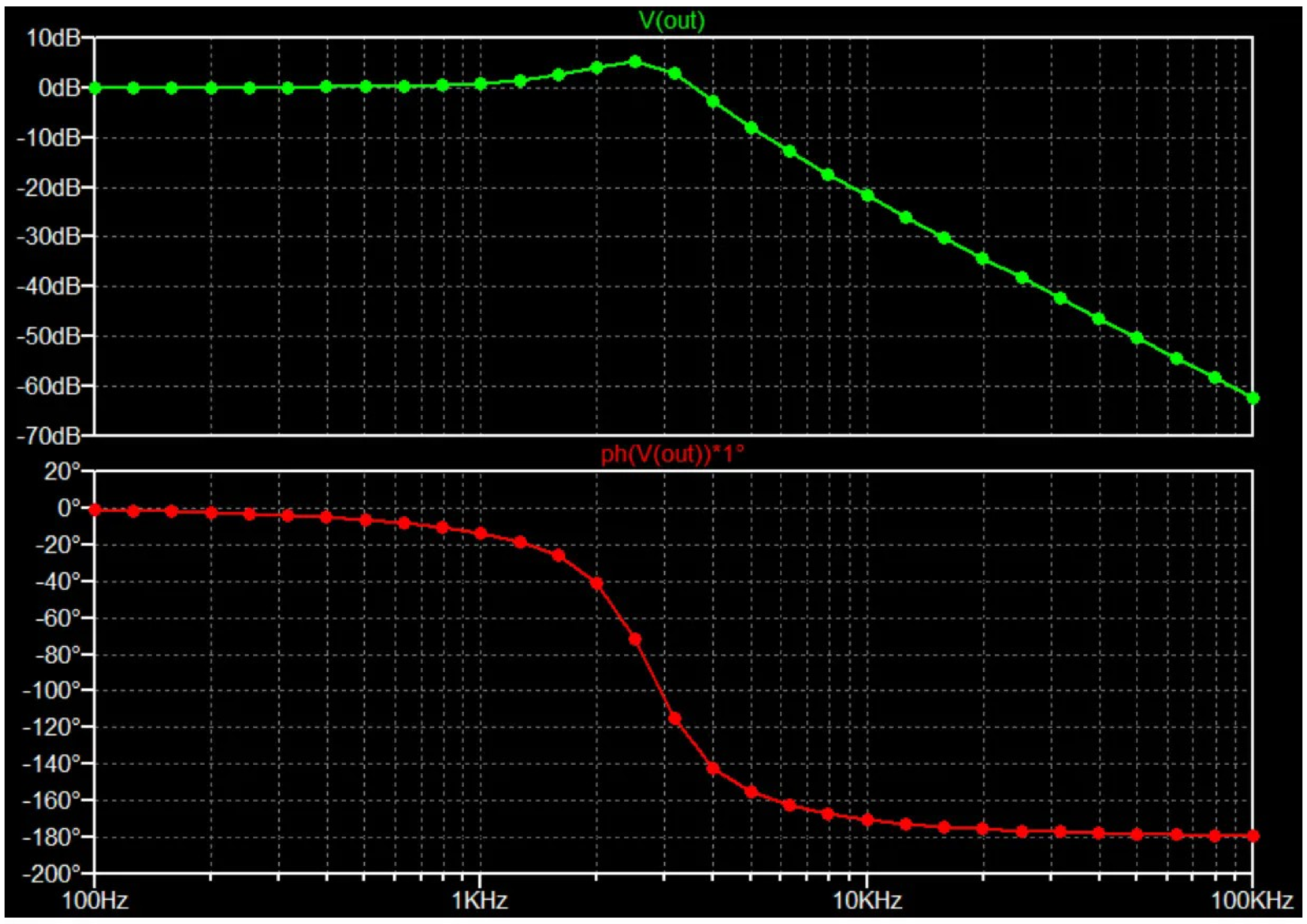


Fig.2: an example Bode plot produced by LTspice for the Fig.3 circuit. The upper plot is gain magnitude (frequency response), with the corresponding phase shift below in red.

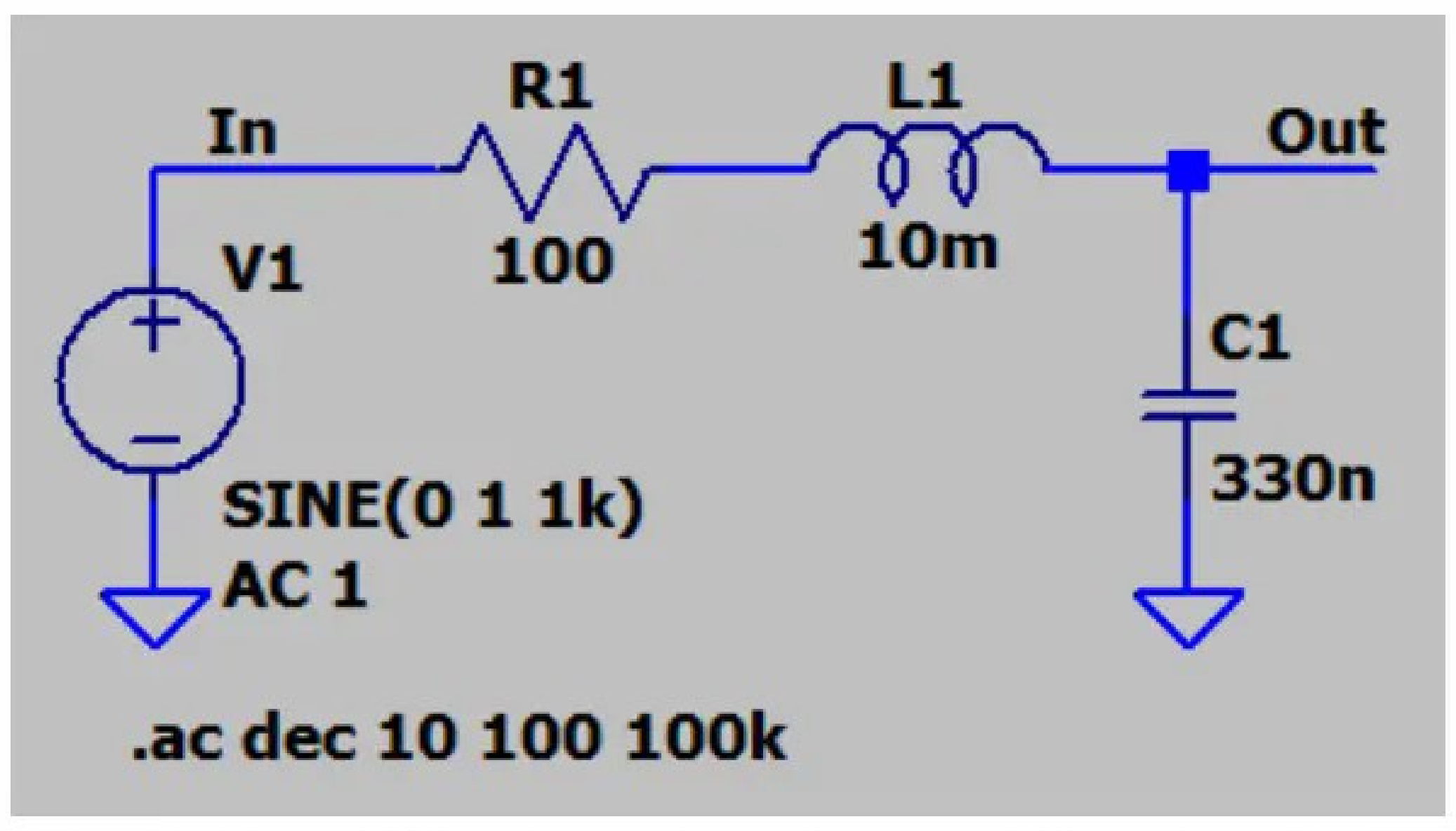


Fig.3: the LTspice circuit used to obtain the Bode plot in Fig.2.

measurements) by applying a range of sinusoidal signals at the input and observing the output.

If we feed a sinusoidal signal to a linear circuit, the output will also be a sinusoid at the same frequency – this must be so because a linear input-output relationship will not change the shape of the waveform. For something other than a sinewave, a different linear input-output relationship may apply at different frequencies.

This means that the component sinewaves that build the waveform may not have the same amplitude or phase relationships relative to one another at the output. The waveshape will change if this happens, but the frequencies present will still be the same as in the input.

This does not mean that it is impossible to measure frequency responses with inputs other than sinusoidal signals, but it makes things more complex, as we have to find the responses at individual frequencies from the combined response. This can be done using analysis based on the fast Fourier transform (FFT), which takes a waveform in time and finds its frequency content (or another type of Fourier transform).

For a linear circuit with a sinusoidal input, the output waveform may be shifted in time, and may have a different amplitude (see Fig.1). When considering frequency responses, we usually represent the time shift as a phase shift, ie, the shift as a proportion of the cycle time (1/*f*) of the waveform at the frequency of interest (*f*) expressed in degrees (the full cycle time is 360°).

For example, at 1kHz, a shift of 0.25ms relative to the input is a quarter of the cycle time of 1ms, so it's a phase shift of 90° (360° ÷ 4).

The basic manual approach to measuring a frequency response is to display the input and output waveforms on a dual-channel oscilloscope, and measure the amplitudes and phase shift, as shown in Fig.1.

The gain of the circuit is calculated by dividing the output amplitude by the input amplitude. The phase shift is obtained by measuring the time difference between equivalent points on the two waveforms and converting this to the phase value.

Whatever method is used to produce the data, the most common way of presenting frequency response data (simulated or measured) for electronic circuits is the Bode plot: graphs of gain in decibels and phase shift in degrees against a common logarithmic frequency axis (see Fig.2). The gain in decibels is $20\log_{10}(V_{out} \div V_{in})$ for voltage signals.

For completeness, the Bode plot in Fig.2 is for the RLC circuit in Fig.3, but the point of Fig.2 is to show the typical presentation of a Bode plot, not to discuss details of the circuit.

The Bode plot is named after Hendrik Wade Bode (1905-1982). His work on this, dating to the 1930s, related to obtaining straight-line asymptotic approximations of the gain and phase responses of systems to assist in calculating their stability. Today, the term is widely used to apply to fully detailed frequency response curves, not just the straight-line approximations.

Phase and steady state

A phase shift may be positive or negative, with a negative phase shift implying the output waveform changes after the input (sometimes called phase lag). A positive shift is the opposite – the output changes occur before the input (phase lead).

A positive phase shift may seem impossible – it appears to indicate the circuit is predicting the future input to produce the output. If this actually happened, we would have a non-causal circuit, which cannot be physically implemented for real-time operation. However, a simple high-pass filter using a single capacitor and resistor has a positive phase shift.

The solution to this potential dilemma is that frequency response is defined for a steady state, that is, under conditions where the input has been applied for a long time (ideally for all time) and the circuit has settled into a stable input and output waveform. Under these conditions, the output can have any positive offset in phase without the system needing to be clairvoyant.

If we apply a sudden input change (a step, or just switching on a sinusoid), the system will not be in a steady state

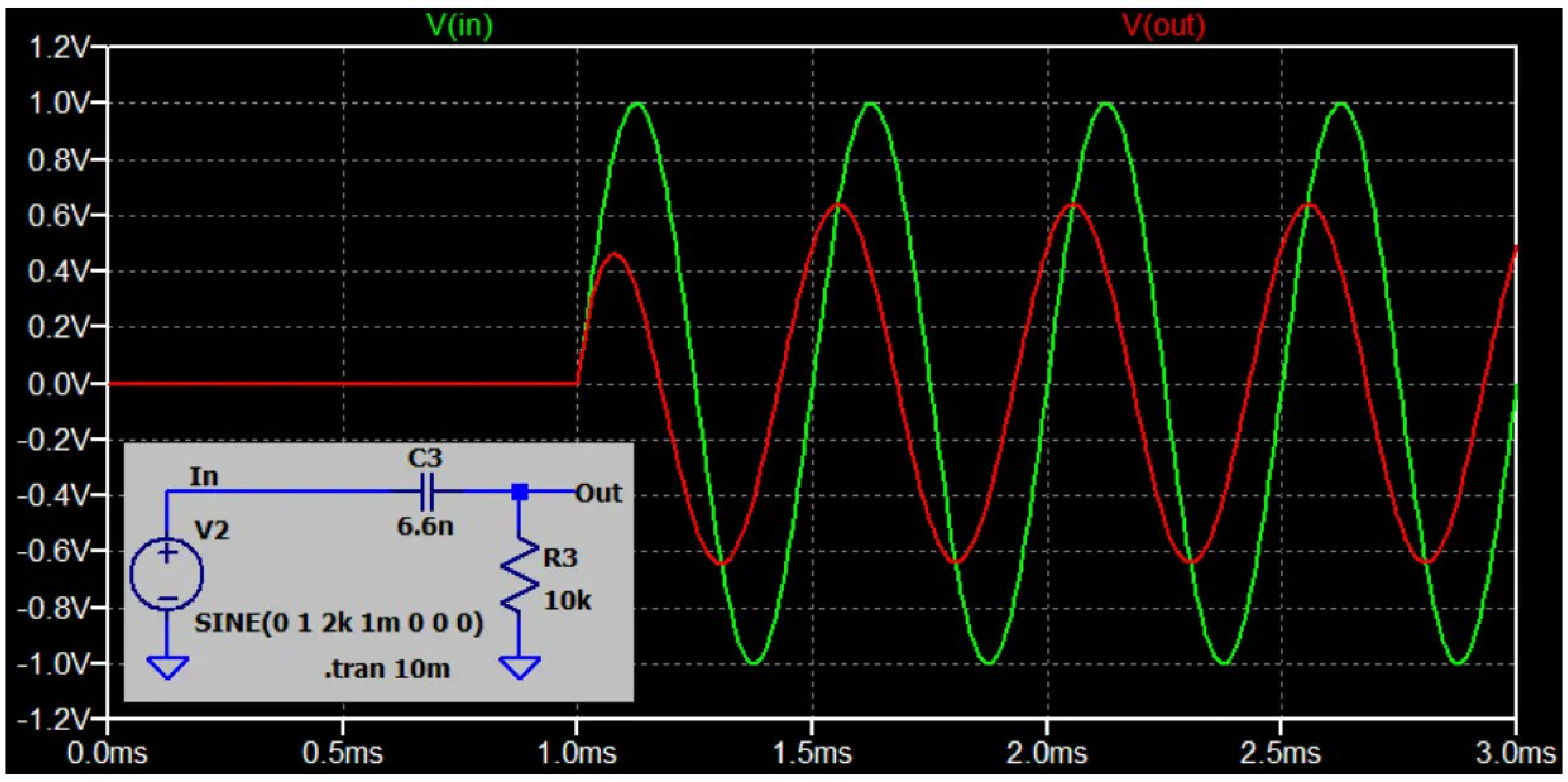


Fig.4: the response of an RC circuit when a sinusoidal input signal switches on abruptly.

directly after the change. For example, Fig.4 shows a sinusoid applied to an RC high-pass filter just after switch-on.

During the first half-cycle, the output (red trace) does not lead the input, but the circuit settles to a state where the output has a positive phase shift, similar to the plot on the right of Fig.1.

A steady-state sinusoid is a requirement for there to be a single frequency at the input. In theory, steady state requires infinite time to achieve, but in practice, most circuits will reach a point that is close enough in a relatively short time. However, the time required for it to stabilise may still need to be taken into account when making measurements.

The theoretical need for a steady state for a perfect sinusoidal input and output does not mean that we have to try to achieve this to make a frequency response measurement. It only applies to the case where we directly measure the amplitudes and phase shifts at individual frequencies (as in Fig.1).

As indicated above, with appropriate mathematical processing of the signals, other input waveforms can be used. We will discuss this in more detail later.

Complex numbers

A key thing that Fig.1 shows is that there are two changes occurring from input to output – the phase shift and the amplitude change. If we write $V_{out} = GV_{in}$ for an amplifier at the frequency of interest, we can say that if $G = 10$ and the input is 1V, the output will be 10V. However, this does not tell us anything about the phase shift that is also part of the transfer function.

This means that in $Y(f) = H(f)X(f)$, we cannot simply say $H(f) = G$. The terms in $Y(f) = H(f)X(f)$ have to encompass both the amplitude and the phase. Ordinary numbers have only one value, so we need something more.

We can represent the phase and amplitude of a sinusoidal signal with amplitude A and phase Φ directly by writing it as $A\angle\Phi$, which is called the polar form. This is depicted in Fig.5 as a line of length A at an angle of Φ to the axis.

The signal's value is represented by the point at the tip of the arrow, which can also be expressed in terms of its Cartesian coordinates (x and y on a graph), also as shown in Fig.5. From basic trigonometry, $x = A\cos(\Phi)$ and $y = A\sin(\Phi)$.

We could just use x and y as two ordinary numbers, but that is not an effective approach for mathematical manipulation. The way forward is to use complex numbers, such that the value of the signal in Fig.5 is $x + jy$, where j is $\sqrt{-1}$ (for some reason, mathematicians like to use i for this but engineers use j because they use the variable i to represent current).

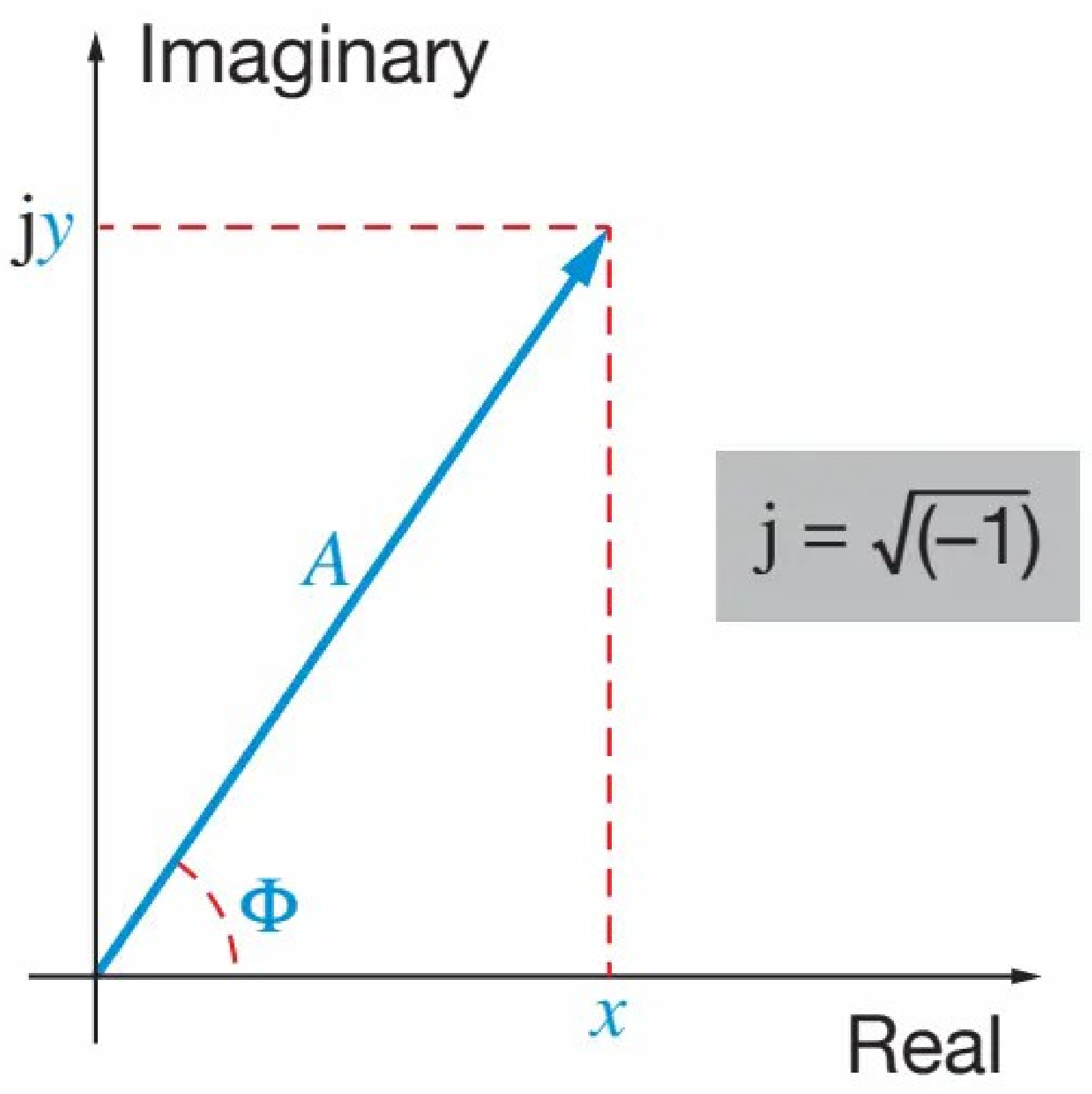


Fig.5: the relationship between phasors and complex numbers.

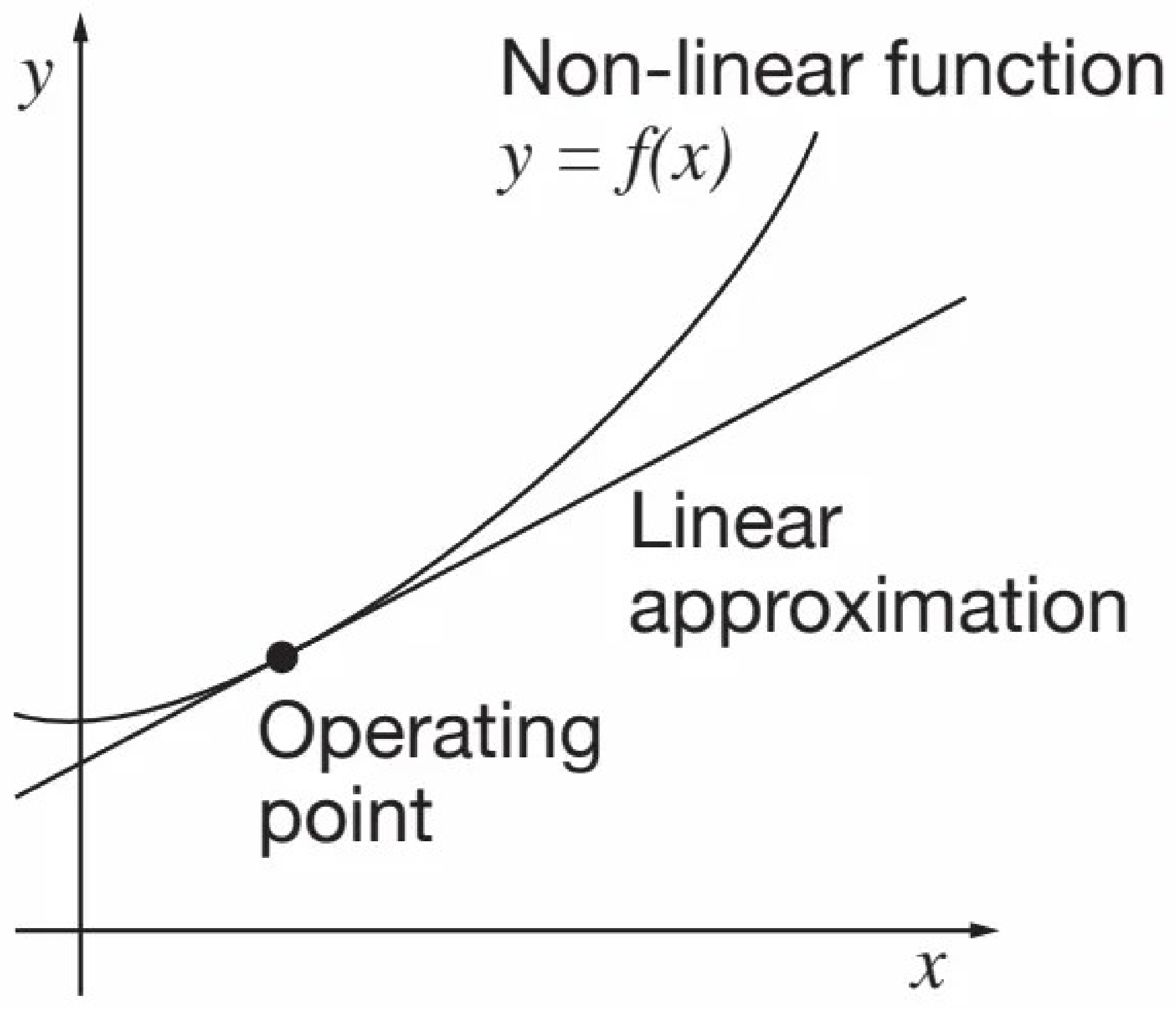


Fig.6: the linearisation of a transfer function at an operating point.

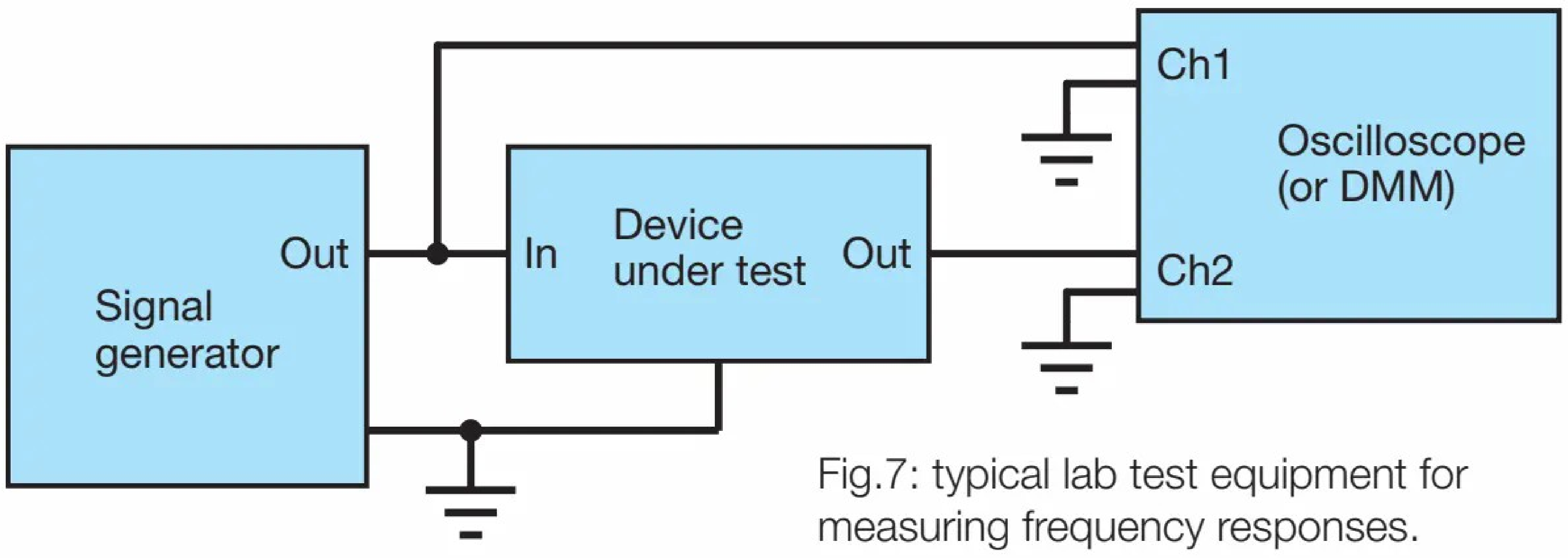


Fig.7: typical lab test equipment for measuring frequency responses.

This is called the rectangular form of the number. In $x + jy$, the x is referred to as the real part and jy as the imaginary part (the square root of minus one does not exist as an ordinary number). Complex number representations of sinusoidal signals are referred to as phasors.

Using complex numbers, two values can be manipulated together in calculations and algebraic manipulations without them interfering with each other (mathematicians call this orthogonality). The use of complex numbers is fundamental to the theory of frequency-dependent impedance and the transfer functions of components and circuits.

You often see transfer functions denoted as $H(j\omega)$ and $H(s)$, where ω is the angular frequency in radians ($\omega = 2\pi f$), and s is the complex frequency variable used in Laplace-transform-based circuit analysis.

The fact that the transfer function is based on complex numbers means that when we refer to just the amplitude part of the gain, we should call it "gain magnitude" or just "magnitude". Bode plots are often labelled this way, but not always. Using the Pythagorean theorem, the magnitude of the signal in Fig.5 is $A = \sqrt{x^2 + y^2}$. Also, $\Phi = \tan^{-1}(y \div x)$.

Circuit analysis based on complex numbers uses advanced mathematics that not everyone will want to study. However, even if it is not for you, it is worth being aware of the basic ideas and notation as you will come across it in IC datasheets, other technical documents, software menus or help documentation. For example, LTspice can display AC analysis results in terms of real and imaginary parts.

Fortunately, you do not need a deep knowledge of advanced mathematics to measure or simulate frequency responses and interpret the results presented as Bode plots. Often, the gain plot will be the main interest – for example, to decide if the gain of an amplifier is flat enough over the frequency range of interest, or if the cut-off performance of a filter is good enough for the application.

However, phase is important too, as it influences signal integrity, distortion and system stability.

An aside on LTspice

An aside for those interested in the finer details of LTspice: you may have noticed the fact that the gain magnitude and phase plots in Fig.2 are in separate plot planes. The default for an AC analysis is to plot both on the same plane, with the phase axis on the right and the phase graph having a dotted line in the same colour as the gain.

When you add the data points, as in Fig.2, the phase line can look quite messy. I do not know of a way of changing the default phase plot to use a solid line, a different colour, or the left-side axis. The traditional form of the Bode plot is to use separate graphs for gain magnitude and phase, but for software tools such as LTspice, it makes better use of screen space to display both together.

Plotting the gain in separate panes in LTspice is simple if you just plot the output twice and switch off one

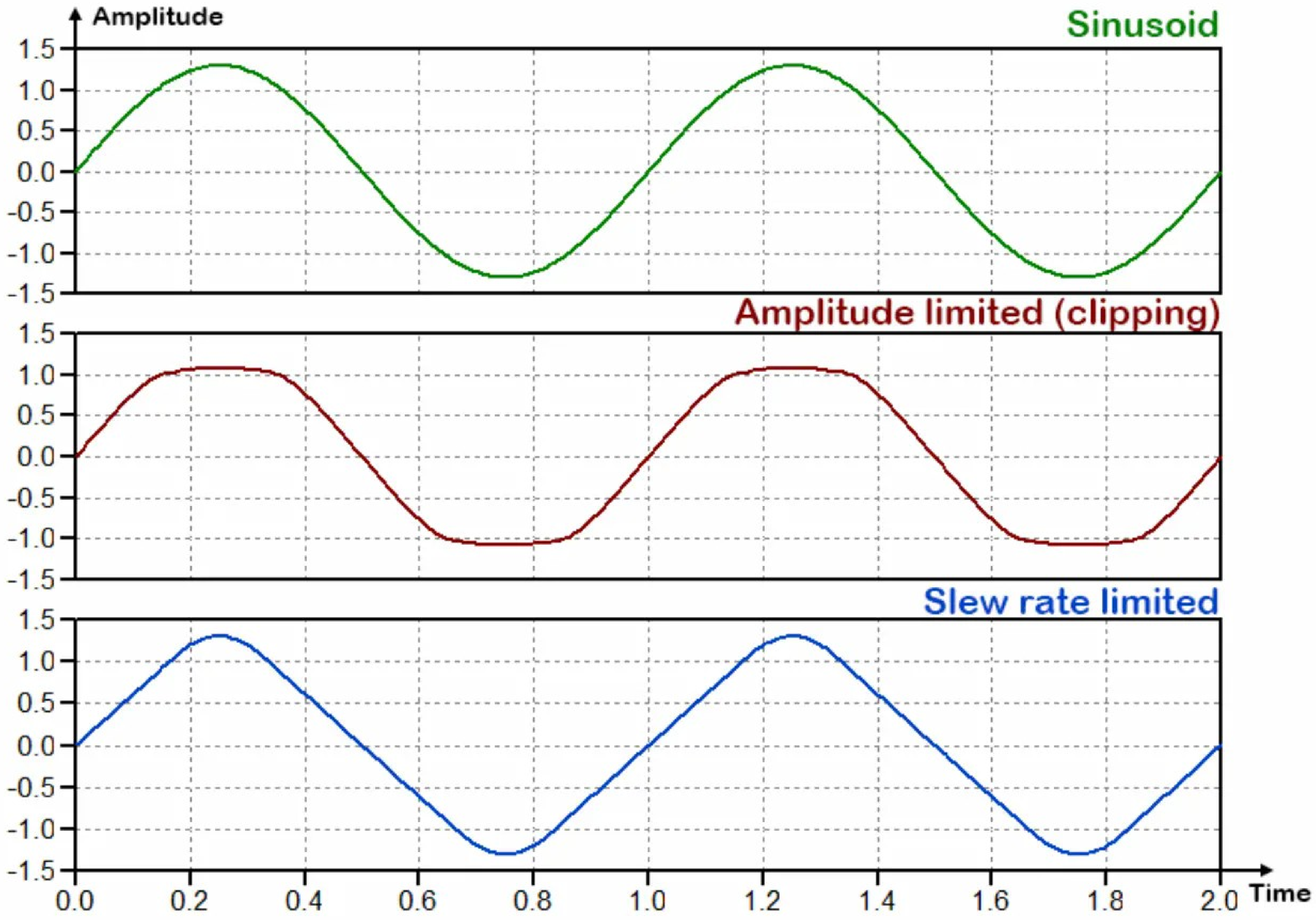


Fig.8: examples of sinusoidal signal distortion. The axis units are arbitrary.

of the plots in each plane off, by right-clicking on the relevant axis. However, the phase still has a dotted line, and the axis on the right. The solution is to plot **`ph(V(out))*1°`** instead (assuming the output signal is named **`out`**).

This uses LTspice's "waveform arithmetic". To do this, you need to click in the pane and "Add Traces", then type in the expression.

Just plotting **`ph(V(out))`** will not show negative phase values correctly, because the trace plotter is interpreting the data as a magnitude, which is what would usually be done for gain in a Bode plot as the default for an AC analysis.

This can be fixed by right-clicking the axis and selecting Cartesian rather than the default Bode representation (polar form). This plots the real and imaginary parts of the data (rectangular form).

The LTspice help on waveform arithmetic tells us that the **`ph(x)`** function will "return a complex number with the real part equal to the phase angle of the argument and the imaginary part equal to zero". We can right-click and switch off the imaginary axis (on the right) – as just stated, all the data points are zero.

The numbers on the left axis are written with "r" (for real number), which is not ideal. Including **`*1°`** in the waveform expression fixes this.

Simulation and measurement

The theory of circuit frequency response is based on an assumption of linearity, but real circuits are not perfectly linear, so we need to consider the implications of this.

When mathematically analysing circuits such as transistor amplifiers, it is common to use what are called small-signal models to approximate non-linear relationships (eg, $y = f(x)$ in Fig.6) to linear ones at an operating point (specific value of x). As long as the value of x does not deviate much from the operating point (hence, "small" signal) the linear approximation is good.

LTspice (and other SPICE simulators) can simulate the frequency response of circuits using AC analysis (as in Fig.2). This works by calculating the operating point and linearising the current-voltage relationships of nonlinear devices such as transistors. The linear models allow very fast calculation of data points on the frequency response without having to simulate using input waveforms.

AC analysis in SPICE does not work in all situations, such as sampling and switched circuits. LTspice is also able to perform a frequency response analysis (FRA) simulation, which follows a process much closer to what might be used for practical measurements.

It performs transient simulations with sinusoidal waveforms over a range of frequencies and calculates the frequency response from the resulting signals. This is more complicated to set up than AC analysis. We used this approach for a sampling circuit in the DSP series (December 2024) and covered the basics of LTspice FRA in the March and April 2024 issues.

The basic lab setup for frequency response measurements is shown in Fig.7. This can be used for manual or automated measurements. For manual measurements, set the frequency on the signal generator and measure the signal amplitudes and time difference using the oscilloscope. Calculate the gain and phase shift as detailed above with reference to Fig.1. Move on to the next frequency and repeat.

Using a suitable DMM or audio millivoltmeter will require you to switch the connection between input and output (unless you have two), and will only give you gain magnitude values.

For automated measurements, the signal generator may be built into the oscilloscope, or separate instruments could be controlled by a PC or linked by some proprietary system. For radio-frequency work, a vector network analyser (VNA) may be used rather than an oscilloscope, and may also measure other parameters, such as input impedance and reflection coefficient.

In this article, we are focusing on lower (primarily audio) frequencies.

Signals for measurement

Analysis and simulation based on small signals is very useful, but for measuring real circuits, we cannot just apply an arbitrarily small signal because we want the measurement to apply to realistic usage. Furthermore, if the signal levels are too low, noise in the circuit and/or test instruments will reduce the quality of measurements.

For the circuits we are typically interested in for frequency response measurement, such as amplifiers and filters, good circuit design should provide decent linearity over normal operating signal levels. Typically, we want to apply the largest signal that is within the circuit's safe operating range, but which avoids any significant distortion of the output signal.

Common forms of distortion are clipping, where the maximum signal amplitude is exceeded, and slew-rate limiting, where the circuit cannot change its signal level fast enough to reproduce the waveform shape correctly. These are illustrated in Fig.8. Clipping results in sinewaves with flattened peaks, while slew rate limiting makes the waveform more triangular.

Relatively severe cases of distortion can be spotted on an oscilloscope display, but may not be easy to see at lower levels. Some measurement systems will be able to measure distortion directly as well as plot frequency responses.

Many oscilloscopes can display signal spectra, which can be used to help detect distortion. A sinusoidal output should have one peak at the input signal frequency. Other significant peaks, particularly at multiples of the input frequency, indicate distortion. As a sinewave is clipped more and more, it approaches a square wave, which will have high odd (3f, 5f etc) distortion products.

The maximum signal level at different frequencies may be different, and can be determined by different characteristics of the device under test. For manual measurements, we could change the signal level for each measurement, but automated measurements may use a fixed level for all frequencies.

Clipping can occur due to either the input or output level being too high. Exceeding the maximum input level is likely to be the limiting factor where circuits attenuate the signal (large input for small output); for example, filters in the stop-band frequencies. Exceeding the maximum output level is more likely for circuits that amplify in all or part of the frequency range (amplifiers, filters in the passband).

The maximum rate of change of a sinusoidal waveform occurs as it passes through zero. Increasing either amplitude or frequency increases the maximum rate of change, so slew-rate limiting is more likely for relatively large, high-frequency signals. This could occur, for example, for amplifiers and high-pass filters at high frequencies, or low-pass filters close to or just past cutoff.

Clipping problems can also occur in the measurement instrument input, which will invalidate the measurement. Ideally, you will get a warning about over-range/overload or clipping if this occurs.

The above discussion mainly applies to active circuits (eg, using transistors and op amps), but you may want to measure the frequency response of passive circuits (containing a combination of resistors, capacitors and inductors). In general, typical test set-ups are unlikely to exceed component ratings, but check if this may be a possibility.

Also be aware that resonant passive circuits (those containing LC-based filters) may produce voltages greater than the input voltage.

Several types of input signal could be used for frequency response measurement – examples are shown in Fig.9. The most straightforward is a stepped sinewave. As already discussed, this is what is used for manual measurements, but of course, it can be automated too.

Higher-resolution measurements (more frequencies and more accurate response data) will take longer. Time is required for the output to stabilise (approximate steady state) each time the frequency changes.

Alternatives to stepped sines include swept sinusoids (also called chirps), where the frequency varies continuously, or wideband random noise (eg, white noise). These input waveforms contain frequency content at all frequencies of interest, so the output frequency content can be analysed to obtain the frequency response.

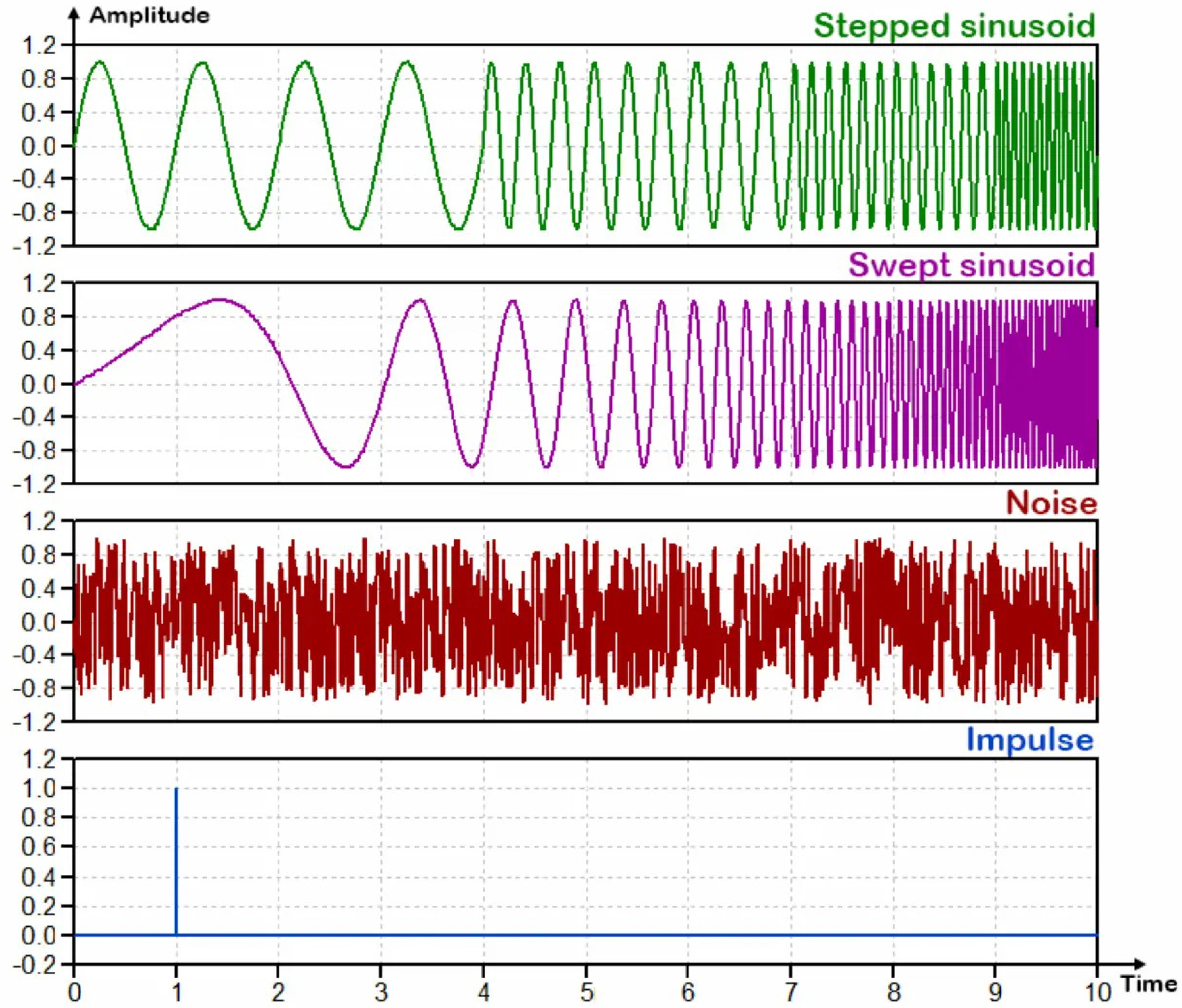


Fig.9: signals that can be used for measuring a frequency response.

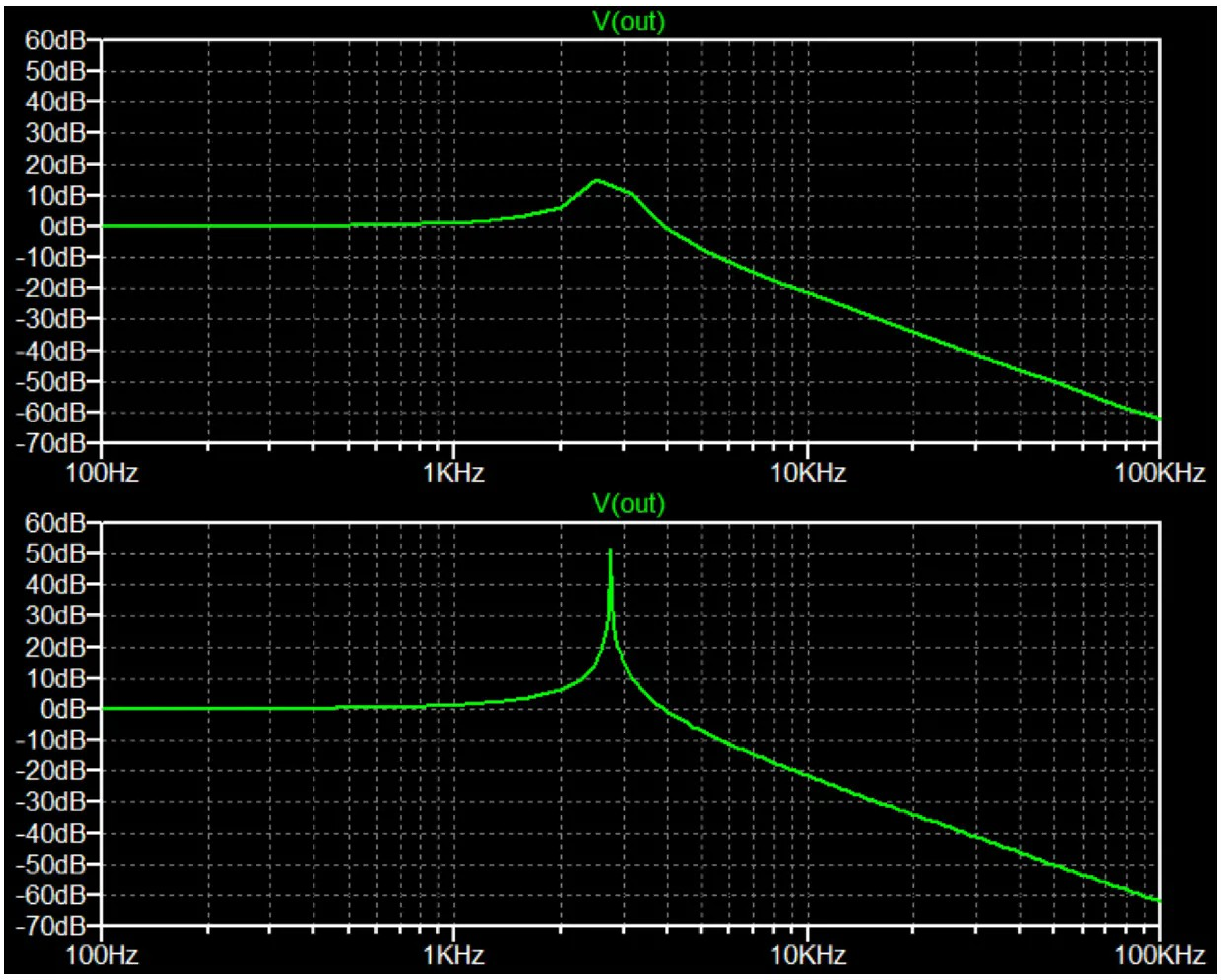


Fig.10: the same frequency response plotted with 10 points per decade (upper) and 500 points per decade (lower).

This is done by processing the waveforms for the whole sweep, or an appropriate duration of noise, using calculations based on the fast Fourier transform (FFT). Logarithmically swept sine measurements tend to be much faster than stepped sine, but the measurements may be worse, particularly at higher frequencies due a lower signal-to-noise ratio (SNR).

In theory, an impulse can be used as the input waveform to obtain the frequency response. Again, an FFT is used to obtain the frequency response. An ideal (analog) impulse is not feasible, but a short high-amplitude pulse will approximate it.

There are practical difficulties – the pulse has to be very short, and therefore causes a small output signal, and hence poor SNR unless it has a very high amplitude, which may not be feasible, or could cause non-linear behaviour. For digital filters, impulses may be usable if a single non-zero input sample can be reliably created.

We discussed impulses in the DSP

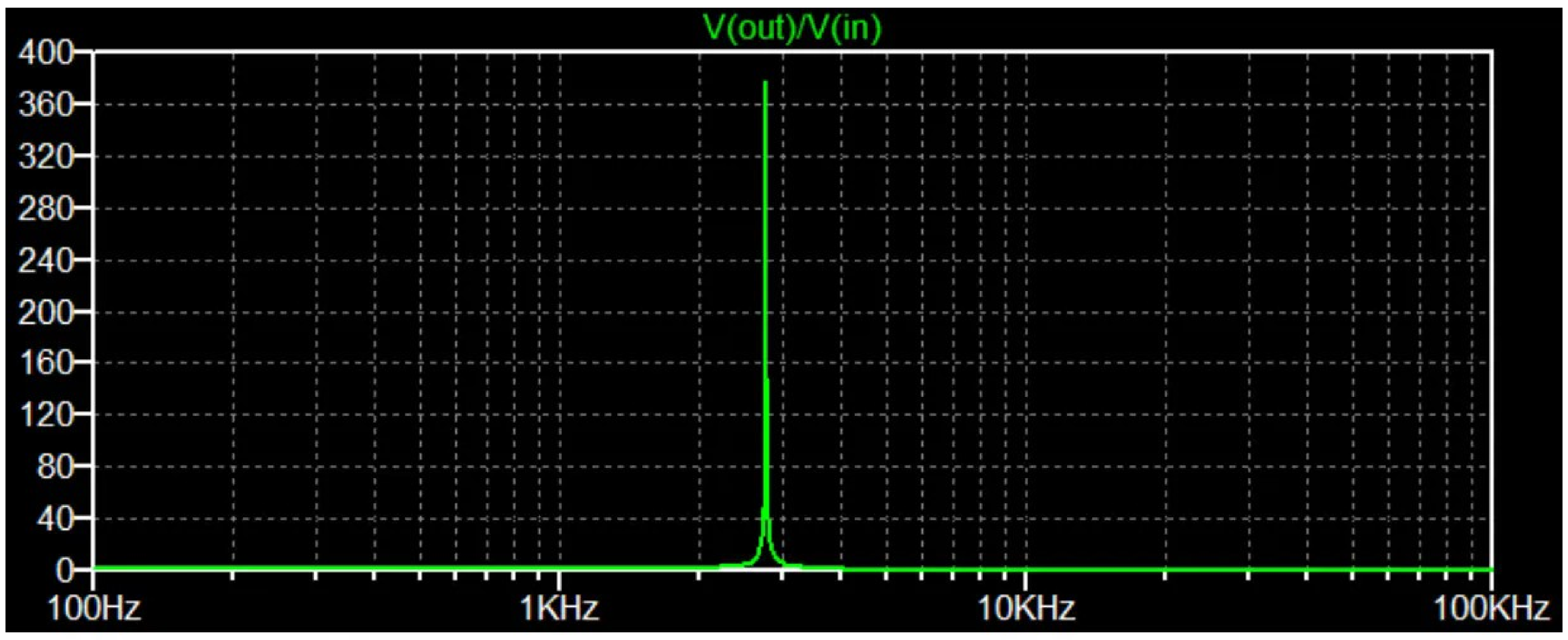


Fig.11: a linear plot of the same response shown at the bottom of Fig.10.

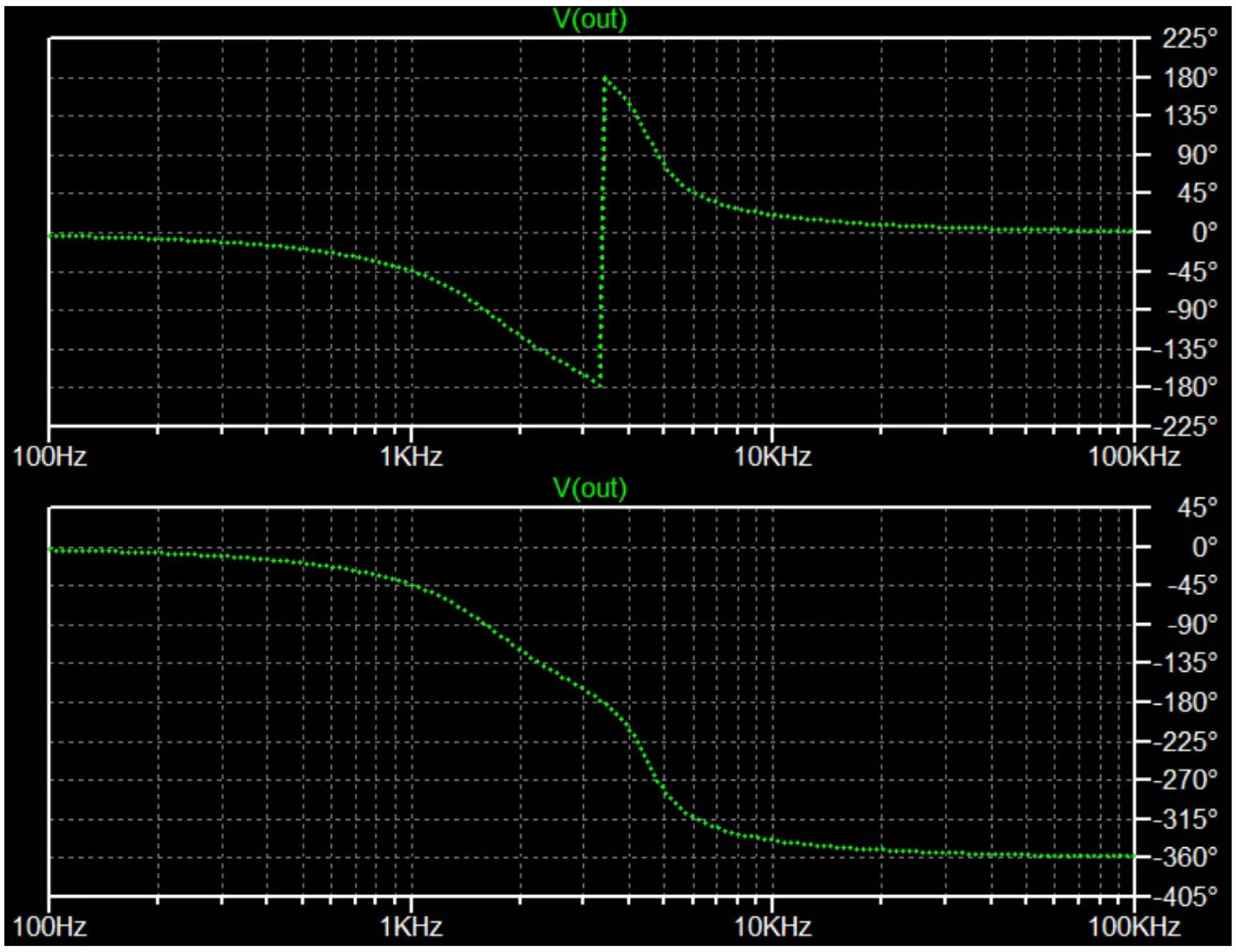


Fig.12: LTspice phase plots of the same network. The upper plot wraps the phase at ±180°, while the lower doesn't.

series, particularly in the August 2024 and April 2025 issues.

Data points and plots

Bode plots usually use a logarithmic frequency axis. This means that frequencies a decade apart (factor of ten) are equally spaced on the axis. For example, there is the same distance from 100Hz to 1kHz on the axis as there is from 1kHz to 10kHz (as in Fig.2).

The number of data points to use is often specified in terms of points per decade, or sometimes points per octave (an octave is a doubling of frequency). Fig.2 has 10 points per decade, shown by the dots on the graph. LTspice's plotting software has interpolated the points between these to draw a smooth curve.

When making manual measurements, how do you choose which frequencies to measure at if you want equally spaced values on a logarithmic plot? Each measurement frequency (f) is a fixed scaling factor (s) larger than the previous one, ie, $f_{n+1} = sf_n$, when n is the measurement number in order of increasing frequency. For p points per decade, if we multiply s by itself p times, we get 10; that is, $s^p = 10$, so $s = \sqrt[p]{10}$.

For example, if we wanted four points per decade, we have $s = \sqrt[4]{10} = 1.7783$. Our frequencies in each decade are the start frequency multiplied by 1, 1.7783, 1.7783^2 and 1.7783^3. For example, starting at 100Hz, we would have 100Hz, 178Hz, 316Hz and 562Hz. Multiplying the final value by 1.7783 gives us 1000Hz – the start of the next decade.

We can calculate all the frequencies to measure by starting at the lowest frequency we need and multiplying by s until we get to the maximum frequency to use.

Similarly, for p points per octave, the scaling is $s = \sqrt[p]{2}$. The number of octaves per decade (N) is $2^N = 10$ (N doublings is equal to ten). So, $N = \log(10) \div \log(2) = 1 \div \log(2) \approx 3.322$.

Insufficient data points can lead to misleading plots. Fig.10 shows the gain response for the circuit in Fig.3 with R1 changed to 0.1Ω. This low resistance results in a very sharp peak, which requires a lot of data points to plot accurately. The upper plot uses 10 points per decade, which is insufficient. The lower plot uses 500 points per decade.

For manual measurements, it can be useful to do an initial run with relatively few data points and then only make closely spaced measurements where the response changes rapidly. For stepped-sine automated measurements, using a lot of data points may take a long time, but at least you can do something else while you wait.

For measurements based on FFTs, the number of data points will depend on the number of waveform samples (in time) and the FFT algorithm. Usually, the number of data points in the frequency response plot will match the number of waveform samples, and will be evenly rather than logarithmically spaced.

Typically, the number of points will be a power of two and be a large number (eg, 16384, 32768 or 65536), so likely to be sufficient for accurate plotting.

Fig.11 shows the same data as the lower plot in Fig.10 plotted using linear rather than decibel gain. We are not really able to see the fact that the circuit is a low-pass filter because the peak dominates the plot. This is a relatively extreme case, but illustrates the usefulness of decibel plots or logarithmic axes in showing details at different scales. Linear axes are more appropriate for plotting smaller ranges.

The symmetry of the sinewave means that just by looking at two waveforms on an oscilloscope, it is impossible to distinguish between phase shifts of +180° and -180° (they both invert the waveform). Similarly, shifts above 180° in one direction look the same as smaller shifts in the opposite direction (eg, +270° and -90°).

Simulations and more advanced measurements may be able to correctly obtain phase values beyond ±180°, but it is still common to plot values in the ±180° range (called phase wrapping).

Simulation and measurement may give you the option to plot phase wrapped to ±180° or unwrapped to the full value; for example, see the LTspice plots in Fig.12, which is for an RLC circuit formed from two copies of the filter in Fig.3. Circuits with relatively large delays may have very large phase shifts at high frequencies, and wrapped plots can show a large number of jumps between ±180°.

Sound card based measurements

There are many PC programs, including some free ones, designed for audio applications such as room acoustics measurement, characterising loudspeakers and other audio devices/systems. These use the PC's sound inputs and outputs to make measurements.

One of their main capabilities is to make measurements to help optimise the acoustics of a room or studio and find the best locations for speakers and listening. This requires a calibrated microphone or sound pressure level (SPL) meter, which may be analog or USB-connected in a setup like in Fig.13.

These programs may provide standalone signal generator and oscilloscope tools, and of course, they can measure and plot frequency responses, as this is a key task in room audio characterisation. They may also be able to measure other parameters, such as signal distortion and component impedance.

Fig.14 shows a simple setup that could be used to test an electronic circuit. This is similar to the arrangement in Fig.7, where the sound card's right (stereo) channel is used as the signal generator output, and the two input channels are used to measure the input and output signals.

Other arrangements may be used; for example, the left I/O channels may be connected together (loopback, as in Fig.13) to facilitate timing measurements. Direct measurement of the input signal may not be needed if the sound card is calibrated first using a direct connection (R out to R in) without the device under test (DUT) present.

Compared with using lab test gear, sound-card-based measurements have limitations. The input and output voltage ranges are limited and may vary significantly, depending on the sound card, particularly between built-in standard sound cards and the external 'professional audio' USB devices often used by musicians.

The frequency range is limited to the audio spectrum (20Hz to 20kHz), although some sound cards with higher ("HD" or "studio") sampling rates (eg, 192kHz) may facilitate higher frequency measurements. Sound cards are designed for audio signals only, and are usually AC coupled, so DC measurements are not usually possible.

Sound cards are not designed as measurement instruments, so are not calibrated. The software usually provides calibration procedures to overcome this. One potential advantage is that they may have greater signal level resolution (eg, 16 or 24 bits) than is typical in oscilloscopes (8 to 12 bits).

The much lower sampling frequencies used by sound cards, compared with what is expected of an oscilloscope, makes this possible.

Measurements may require two inputs (using stereo, as shown in Figs.13 & 14), which should not be a problem on many PCs, but laptops commonly only have a single channel microphone input, often combined with a stereo output via a TRRS (tip-ring-ring-sleeve) headset socket. An external (eg, USB) audio interface can also be used.

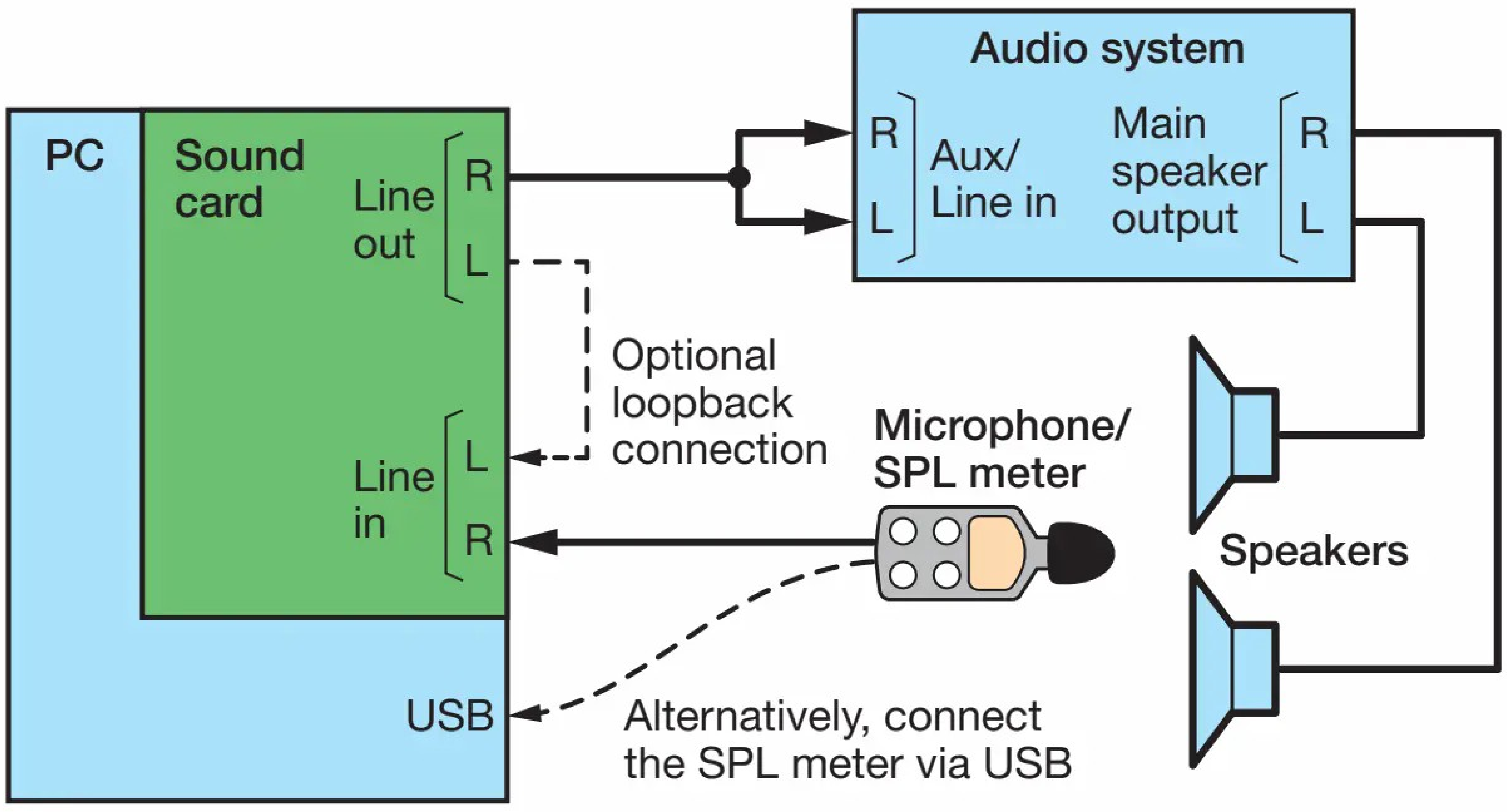
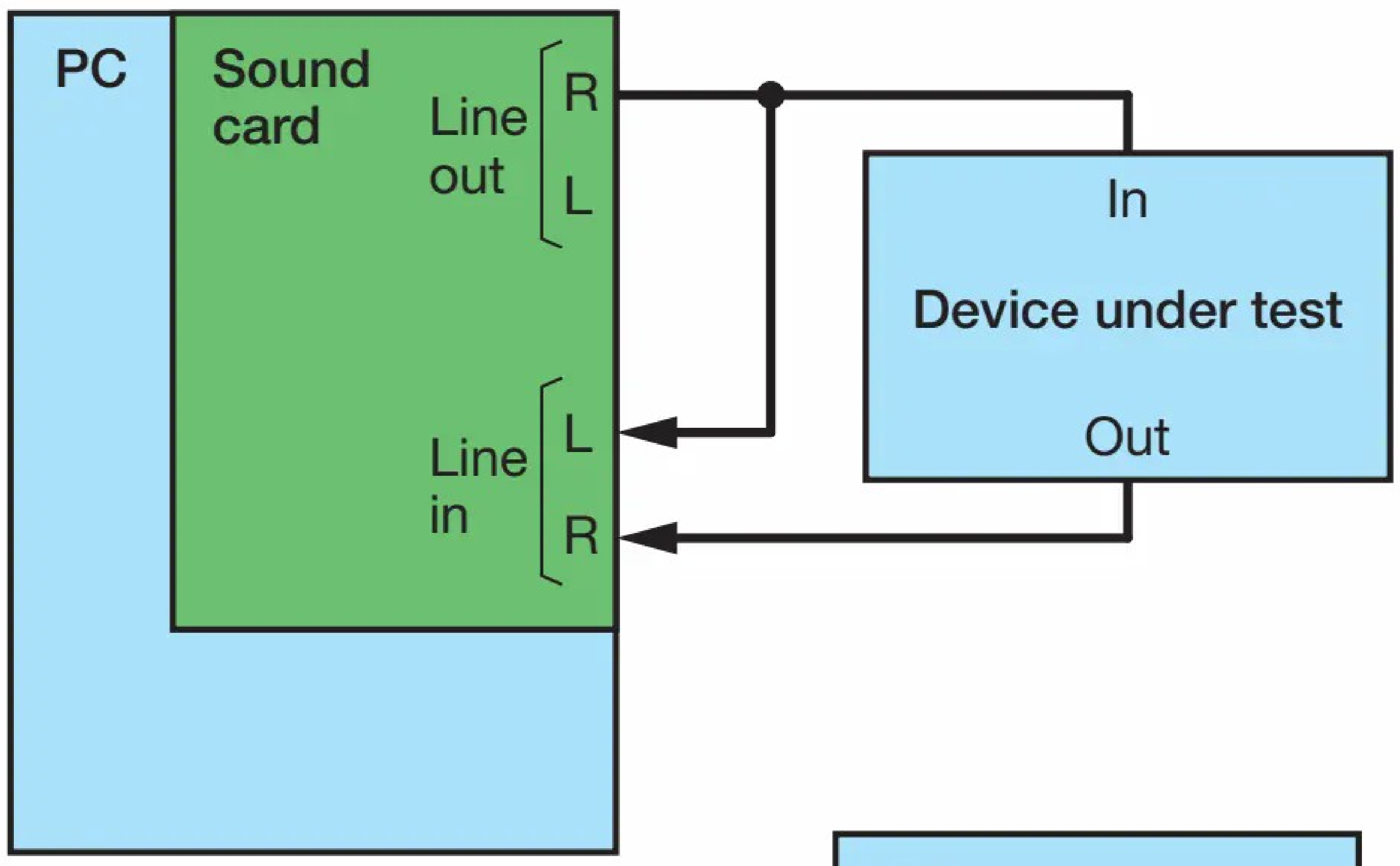

Audio system
PC
Sound card
Line out
R
L
R
L
Aux/ Line in
Main speaker output
R
L
Optional loopback connection
Microphone/ SPL meter
Line in
L
R
Speakers
USB
Alternatively, connect the SPL meter via USB

Fig.13: a typical setup for acoustic frequency response measurement using a sound card. The audio system can be directly measured by attenuating the speaker output signals to levels that are suitable for feeding into the sound card line-in terminals. Only try that if you really know what you are doing!

Fig.14: a simple setup for testing an electronic device using a sound card.


PC
Sound card
Line out
R
L
In
Device under test
Out
Line in
L
R

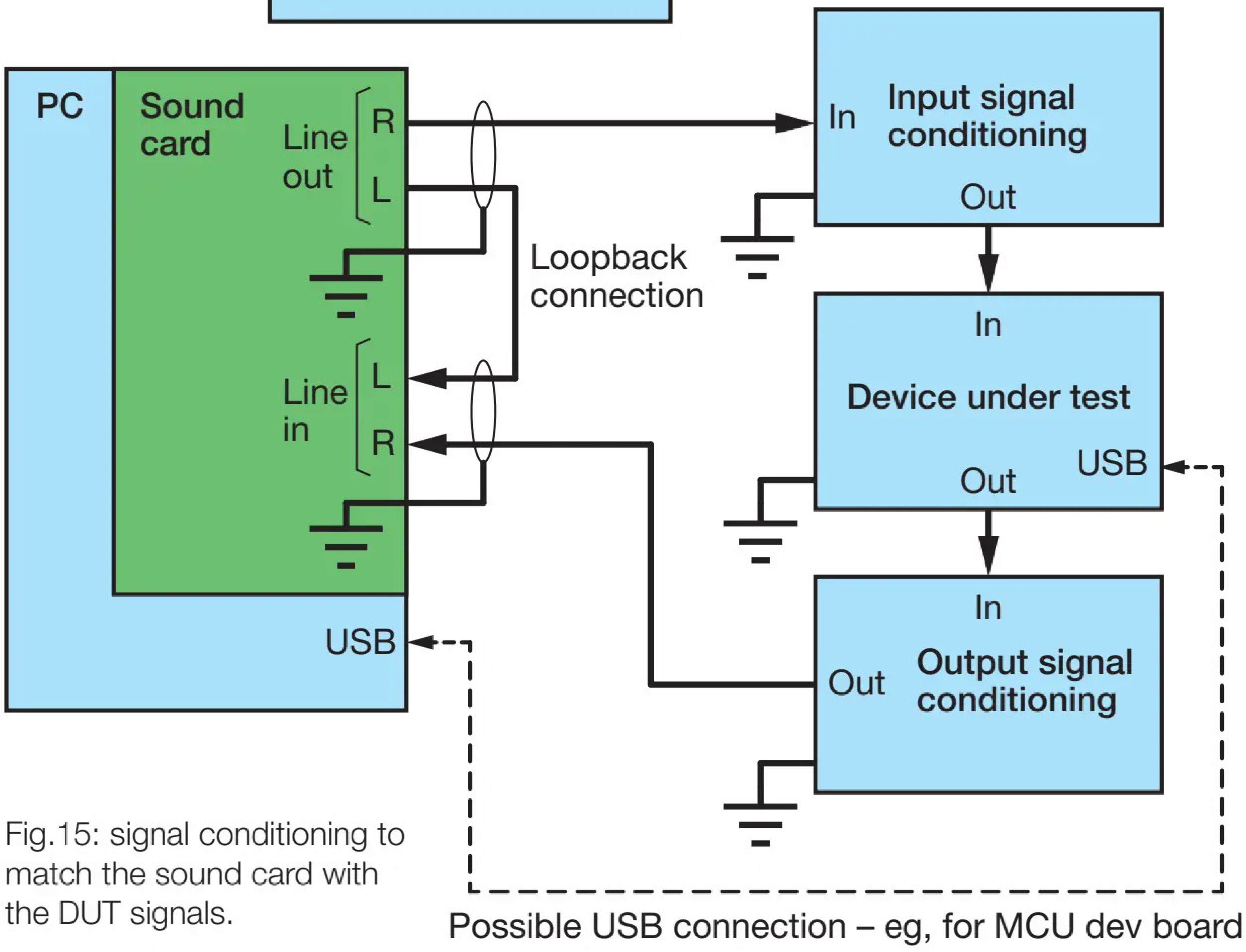

PC
Sound card
Line out
R
L
In
Input signal conditioning
Out
Loopback connection
In
Device under test
Line in
L
R
Out
USB
In
USB
Out
Output signal conditioning
Possible USB connection – eg, for MCU dev board

Fig.15: signal conditioning to match the sound card with the DUT signals.

The arrangement in Fig.14 may work in some cases, but often, the signal levels and drive capabilities for the DUT and sound card I/O may not be compatible. This will require signal conditioning circuits on either or both the input and output of the device (see Fig.15).

Coming up

We will continue to discuss measuring frequency responses with a computer sound card in the second article in this series, published next month.

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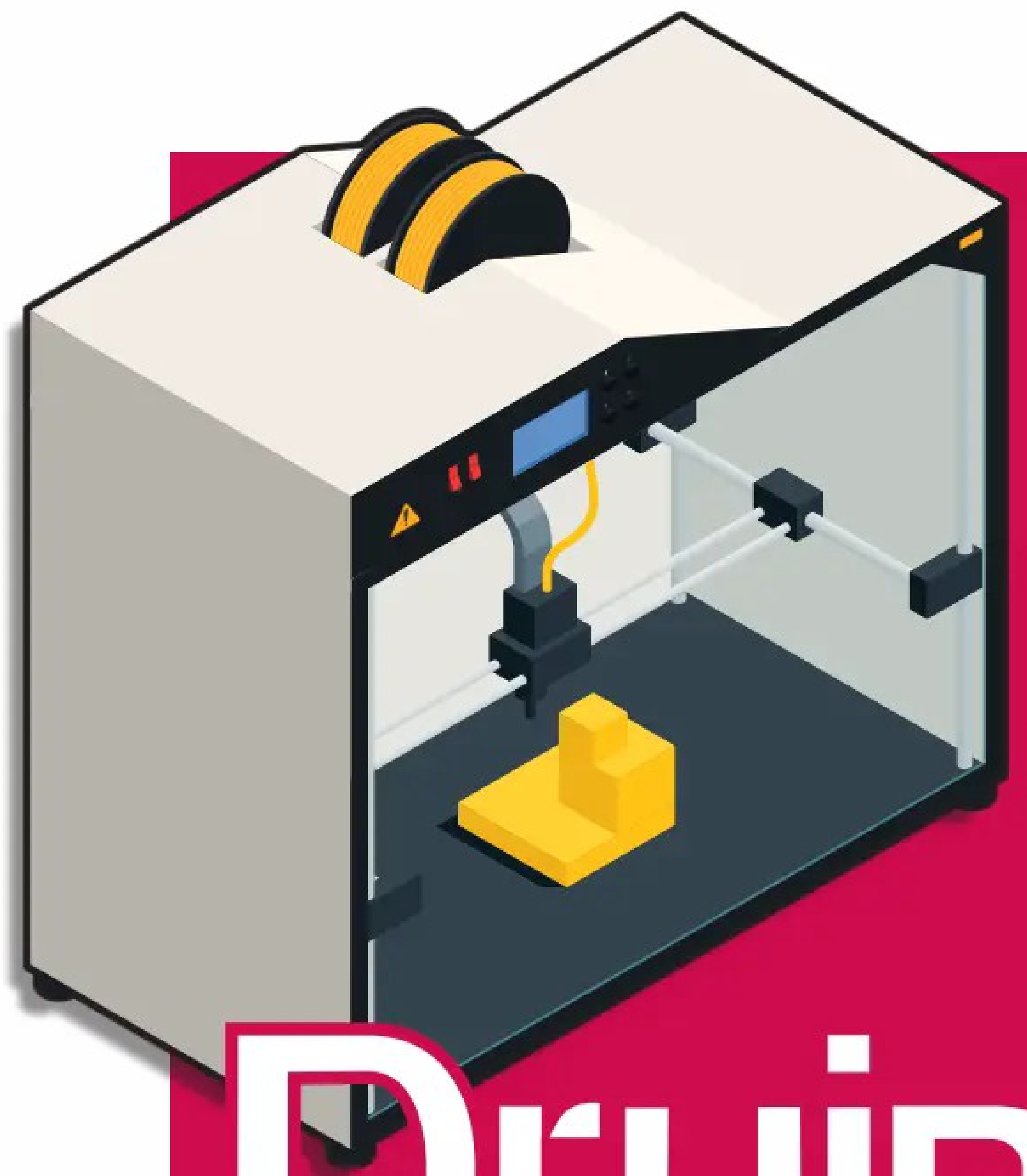


3D Printer Filament Drying Chamber

This enclosure can store up to four 1kg reels of 3D printer filament, keeping them dry and ready for use at any time. You don't even need to remove them – the filament can simply be fed to the printer through a small hole in its lid!

Part 1 by Phil Prosser

The ability to produce functional 3D parts, either standalone or as part of a larger project, is incredibly useful. Over the last few years, 3D printer prices have fallen remarkably. You can now find some amazingly-priced 3D filament printers on the market.

My local electronics retailers stock "Creality" products, which I think are excellent. There are plenty of other good alternatives available online.

My grandson, who wanted to buy printed parts, drew me into this. I pointed out that for the price of a handful of 'bought bits', we could buy our own 3D printer. So I did. I quickly found that being able to manufacture complex 3D parts was incredibly handy.

Like most of these technical things, once you start, there is an amazing range of extras you might want or need. One surprising accessory is a filament dryer. It had not dawned on me that plastic filament can absorb moisture. However, PLA (polylactic acid), probably the most common filament these days, is sufficiently hygroscopic that moisture can become a real problem.

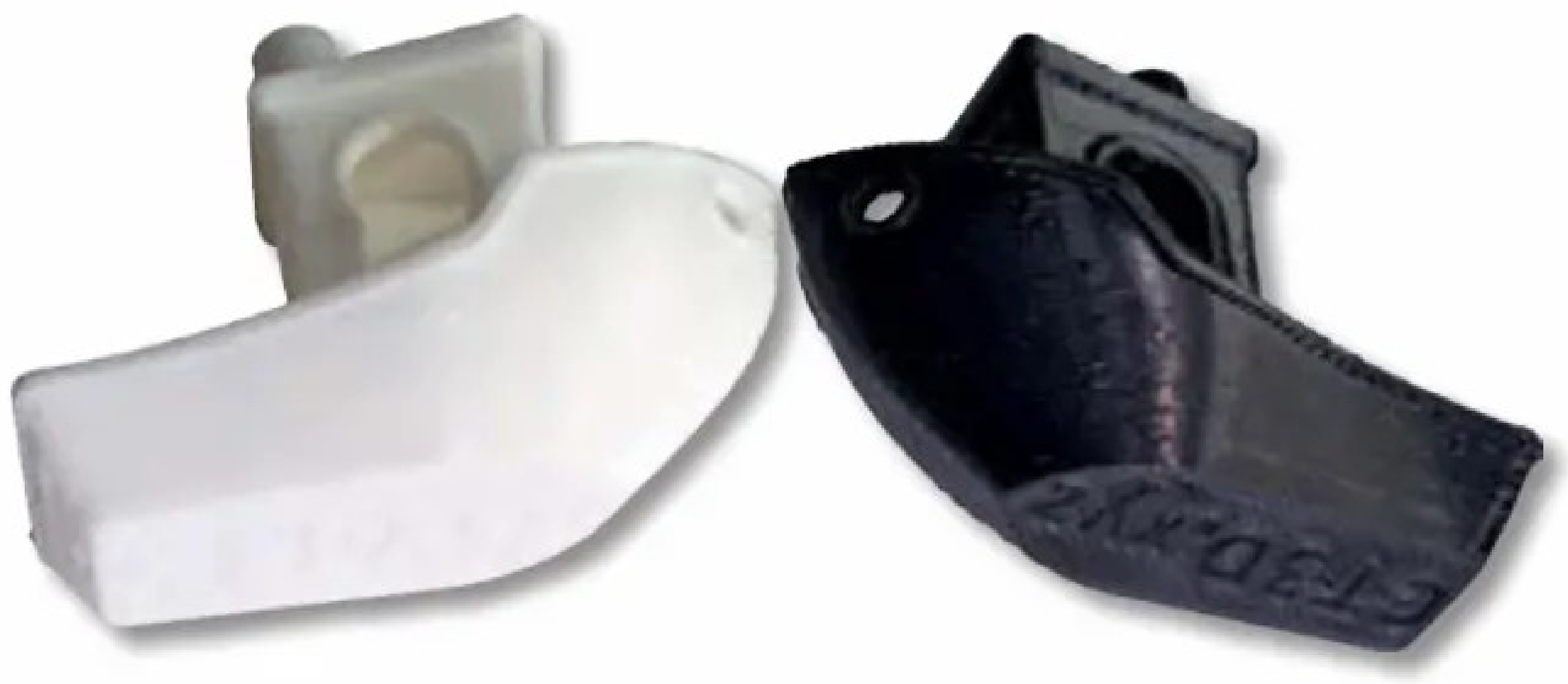


Photo 1: the surface of the black boat is not smooth due to moisture in the filament. The white filament was dry, giving a much better result.

3D printers work by heating the plastic filament to around 200°C (or much hotter for materials like ABS) and extruding it through a small nozzle, typically 0.4mm in diameter. The printer acts like an X-Y plotter and deposits lines of melted filament where required, in layers, thus building the part.

It is incredible to consider that a large print may have the printer laying down material in this manner for 12-24 hours, all without error.

If that sounds too complicated to be reliable, well, you need to get many things right for the printer to work well. However, when set up correctly, reliable results can be achieved. I would say that most electronics hobbyists would have the inclination, skill and inquisitiveness to learn the tricks and tips required to keep a 3D printer running, but they certainly are not 'set and forget'.

When I first ran the printer, things went swimmingly well. However, I later realised that even a little moisture in the filament can cause problems when it is heated in the extruder. The moisture boils into steam, which pushes filament out of the extruder and causes 'blobs' on the print.

Photo 1 tries to show the difference between fresh new filament (white) and some that had been lying around (black).

All the printer knows is that it has driven the correct length of filament at the right time, but the 'blobs' mean it doesn't end up exactly where it should be. So surfaces can get 'blobby', and you hear small popping noises during printing.

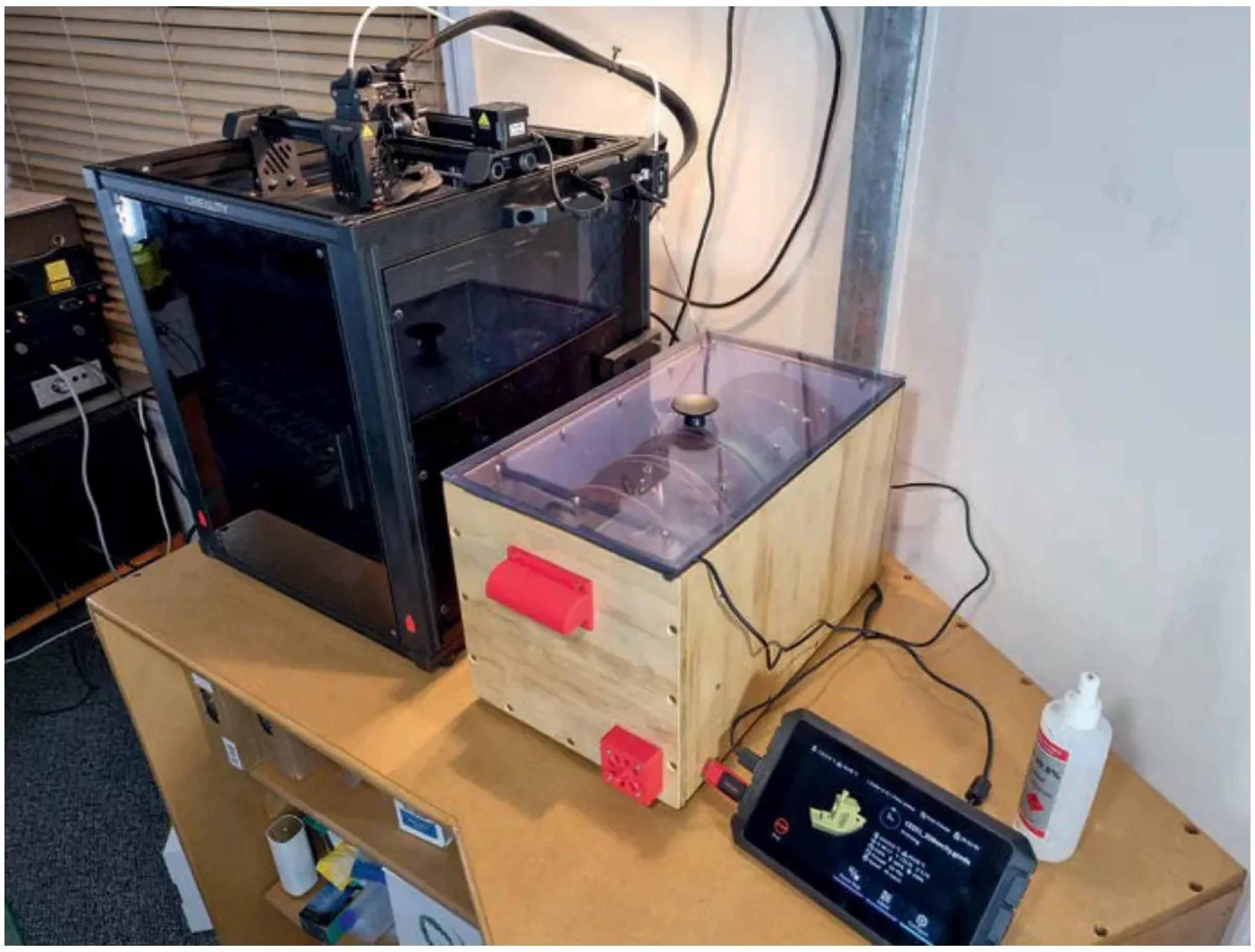
While PLA certainly suffers from these problems, other materials, such as Nylon, also have a terrible reputation for being hygroscopic and hard to print with.

While the printers themselves are competitively priced, I was not really into spending hundreds more on a fancy filament dryer. Some people use a food dehydrator, which, while cheap, does not handle multiple reels or allow you to feed straight from the dryer to your printer.

I was convinced that I could easily make something to do the job with a handful of bits from the spares box, a leftover laptop power supply and maybe a microcontroller. We can even customise the size and shape to suit our workspace and needs. So, while we provide a complete parts list here, you can modify the design to reuse bits you already have, saving a few bucks.

There does not seem to be a specific 'right way' to dry or, perhaps more correctly, dehydrate filament. All approaches use an elevated temperature and some form of timer. Some add air circulation, while a few incorporate

3D Printer Filament Dryer pt1



a mechanism to change the air in the box periodically.

The idea of heating the filament in a sealed enclosure is that when the air in the enclosure gets hotter, it can hold a lot more moisture, so relatively speaking, the air is dryer. In other words, the relative humidity of the air in the box reduces as it is heated.

Fig.1 shows that for a typical room at 20°C and 40% relative humidity (RH), there is about 6g of water per kilogram of air. If the box is sealed, there is always the same amount of water in the box. So, at 42°C, we see the relative humidity will be about 10%. Because the air is now quite dry (for its temperature), it pulls moisture from everything in the box.

PLA filament that has absorbed moisture does not dry out quickly; drying times are typically 6-9 hours. By keeping the dryer sealed and including some desiccant, such as silica gel, in the enclosure, we can keep the filament dry and ready for use. If you will not use the dry filament for a while, it remains a good idea to seal it in a vacuum bag.

The design

This project has two distinct parts. The first is a filament dryer controller board. This is a standalone thermostat controller board that could equally be used to control an incubator or curing oven for painted parts. The board is essentially a thermostat with a timer and substantial protection circuitry.

The second part is making an enclosure for the filament. There are several possible approaches, ranging from very simple to quite complicated. Choosing your approach to the container is probably the most critical choice, as the controller is not that complicated.

We built two enclosures. The first was a custom one optimised for our needs and just a little bit fancy – see Photo 2 and the image above. The second was an 18L plastic tub into which we installed the controller and heater (Photo 3). The latter proved to be quick and simple to assemble and quite effective. It must be said that it looks a lot like a plastic tub, though.

We will provide an overview of how to build the custom enclosure but will not go into great detail. If you are not confident in filling in the details yourself, stick to using the off-the-shelf plastic tub.

Both enclosures use the exact same controller, but we have arranged the heating plates quite differently to suit the differing enclosure shapes. In both cases, we found that without adding insulation to the enclosure walls, we could achieve about 47°C inside with 50W of heating.

Adding a layer of Corflute to the bottom and walls of the enclosures increased the temperature at that power level by well over 5°C, effectively reducing the amount of power needed to keep the enclosure at a given temperature.

The unit is powered by either a 24V DC 4A plugpack or an 18-24V 3A+ DC laptop power supply. Is it just me who has a growing collection of these things, which seem to outlast the laptops they powered? Either way, it drives a resistive heater in the box via a control board, much of which is safety circuitry.

We put a couple of small bags of silica gel in the box to absorb any

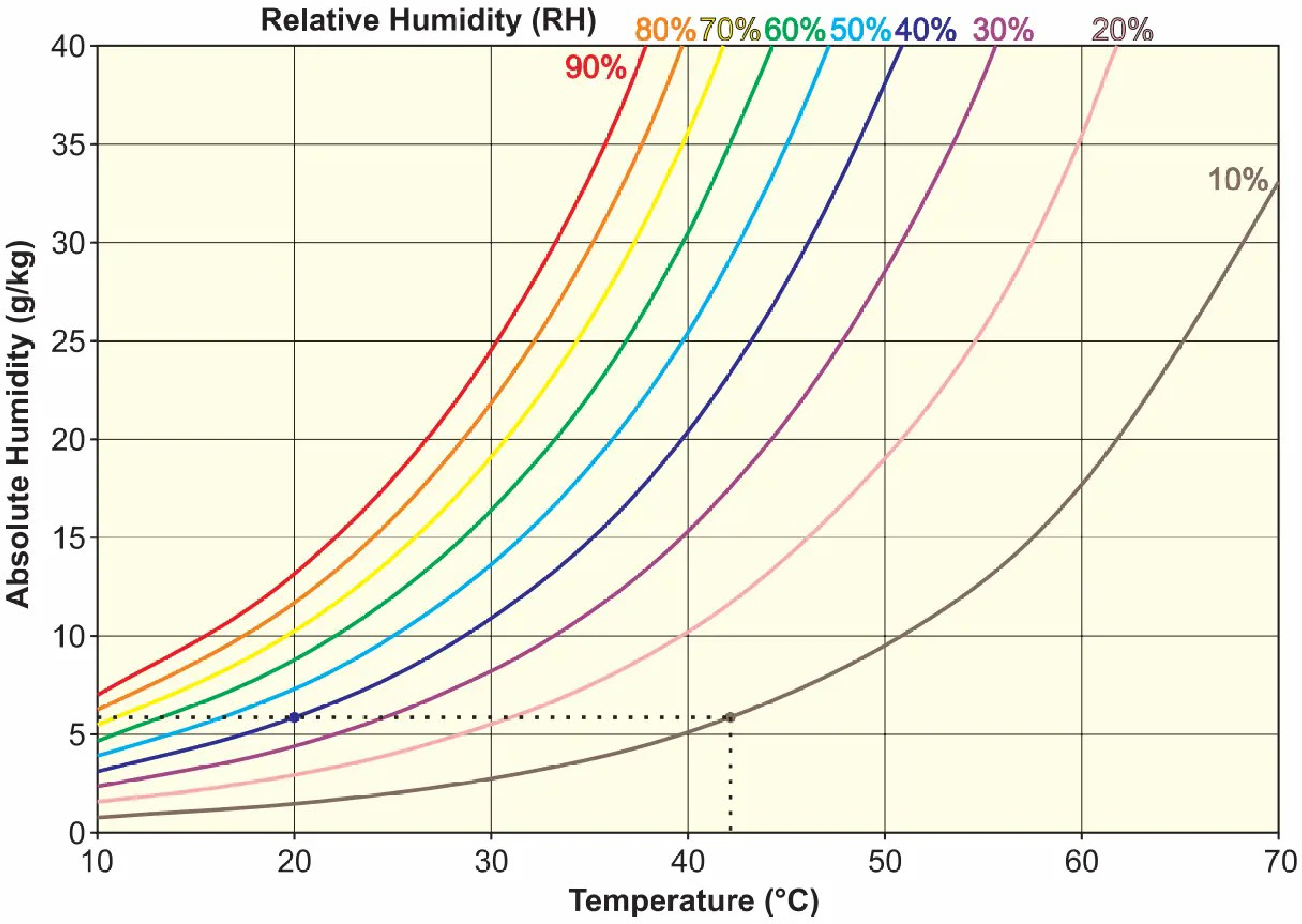


Fig.1: water in the air plotted against temperature for a range of different relative humidity (RH) values, from 10% to 90%. You can see how hotter air can contain a lot more moisture for the same RH figure.

moisture released by the filament and occasionally change the air in the box to expel excess moisture. Cat litter crystals are simply silica gel, so for $10 at the local supermarket, we got a huge bag of silica gel from which we make our own drying sachets. We just put it in paper envelopes to pop in the dryer.

Our filament dryer hangs the reels on a rod and allows you to draw the filament straight from inside the dryer box.

We decided to omit a fancy display, which technically is not hard but adds construction constraints and cost. During development, we noted that even with a fan circulating air in the dryer, the temperature throughout the box varied significantly. So, a temperature display may feel important, but it would only be indicative. Leaving out the display also avoids the need for a humidity sensor.

This decision was hard but it keeps things simple and cheap. If the box is warm and you have fresh silica gel, after a couple of cycles, your filament will be as dry as it will get. Some really cheap humidity sensors are available online that you can pop in the box if you want to monitor it.

Because we are making potentially combustible materials hot, we have taken a very conservative approach to the design to ensure that it is as safe as reasonably possible. Refer to the text box on safety analysis for a discussion of how key design drivers were arrived at. If you are designing your own enclosure, you should consider the hazards we list and satisfy yourself that your approach mitigates all hazards.

The design presented here is mostly about implementing the control and safety systems identified in Table 1, which mandate the following inclusions:

- A controller that maintains the Dryer in a safe state until the user deliberately starts a cycle.
- A thermostat, allowing the temperature to be set from room temperature to 50°C.
- A timer that allows a six- or nine-hour drying period, then shuts the heater down.

Table 1 – Hazard & Risk Assessment

Hazard	Initial Risk	Mitigation	Final risk	*Consequence if not mitigated*
Uncontrolled heating, causing the enclosure to become excessively hot	High	Implement a temperature control system. Limit the maximum energy available so the ultimate temperature without control is safe (50W gives a maximum of around 60°C).	Low	*Damage or combustion of filament or enclosure*
Short circuit or critical component failure	Low	Integrate thermal switches/fuses that disable the system at a safe temperature. Include a fuse in the design, to blow in case of a catastrophic short.	Low	
Excessive heating since the control system does not sense the real temperature	Moderate	Include a fan to circulate air throughout the enclosure.	Low	
Failure of fan results in loss of thermal control	Low	Integrate a 'fan operating' sensor and shut the heater down if the fan fails.	Low	
Heating element contacts personnel	Medium	Mount heating resistors inside a plenum or behind sheet aluminium to minimise the likelihood of contact with personnel.	Low	*Personal injury*
User touches energised part	Medium	Operate the dryer from an isolated plugpack with a low voltage output.	Low	*Electric Shock*
Long-term heating results in auto-ignition of material	Low	A timer shuts the unit down after six or nine hours	Low	*Fire and uncontrolled energy*
Enclosure operates unexpectedly	Medium	The system starts in an idle state. Force the user to press a start button to commence drying.	Low	*Inadvertent operation*
Software fails	Low	Critical controls (thermal- and energy-related) are to be implemented in hardware.	Low	*Inadvertent operation*
Heating element touching combustible material	Medium	Limit the heating power such that the element does not exceed 80°C. Mount the heating element so it is not in permanent contact with timber. Use polypropylene Corflute for insulation, which has an autoignition temperature of 288°C (flash point 260°C).	Low	*Fire*
Misuse – user fills the enclosure with rags or paper	Medium	Integrate thermal cutout on heater plates at 90°C (high but safe).	Low	*Fire*
Misuse – user covers the dryer with a blanket	Medium	Use a thermostat to control the internal temperature, with a safety shutdown & timer.	Low	*Fire*

- Onboard fusing.
- A thermal cutout on each heater element.
- A thermal fuse on the controller board.
- The maximum heating power is limited to 50W.
- A ventilation fan that is integral to the controller board, ensuring airflow in the box.
- An interlock that shuts down the heater if the ventilation fan stops.

We have spread the heating across six 25W resistors, which dissipate 8W each into the large aluminium heating element. Even if everything fails, they will never get hot enough to create a hazard. We tested our two boxes with all controls disabled and determined that 50W of heating resulted in a maximum box temperature of no more than 60°C.

Looking at what is on the market and having read a lot of tests on commercial filament dryers, most make wild claims as to the temperatures they achieve. We feel that 50-55°C is a good, safe temperature. If you want it to get hotter, you would need to increase the power or reduce the size of the box. The controller will accommodate that, but we advise you approach any changes with appropriate caution.

You may have your own spin on how to build this; you could design a box that better suits your needs and use a surplus power supply. You could even reuse some different heating resistors. That will let you build a dryer for a fraction of the cost of a ‘bought one’, but make sure you follow our safety tips so everything goes well for you.

We will first describe the controller and then present a couple of ways it can be used.

The controller

The controller can operate from 18-24V DC, so you can recycle a laptop supply or similar power brick. It must deliver sufficient current for your resistor bank. The input is fused; select a fuse rating an amp or so above your expected maximum operating current.

There is also a polarity protection diode that will dissipate about 2W; we have included heatsinking fills on the PCB, and this ‘extra power’ simply adds to the overall heating in the system.

The controller is expected to be

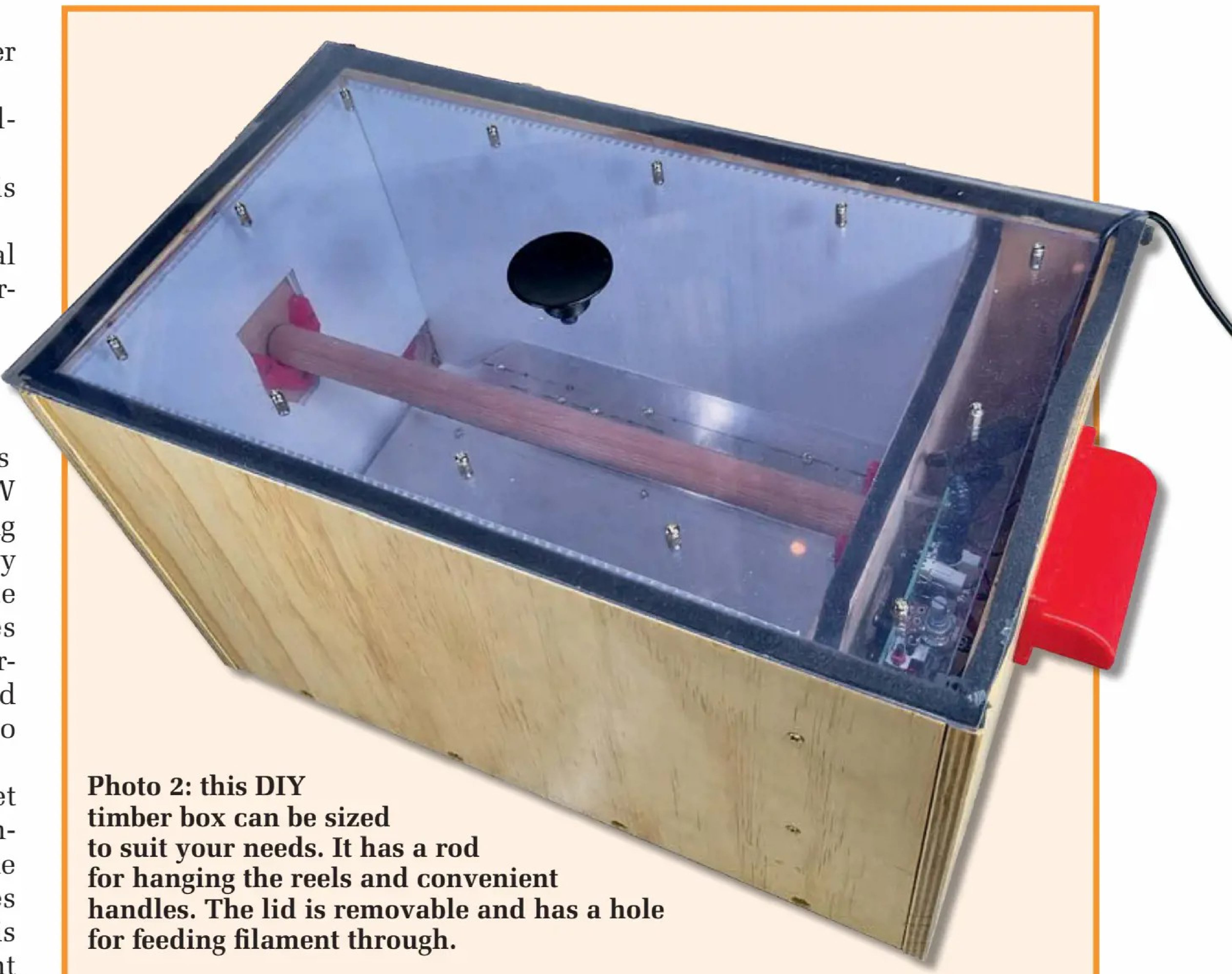


Photo 2: this DIY timber box can be sized to suit your needs. It has a rod for hanging the reels and convenient handles. The lid is removable and has a hole for feeding filament through.

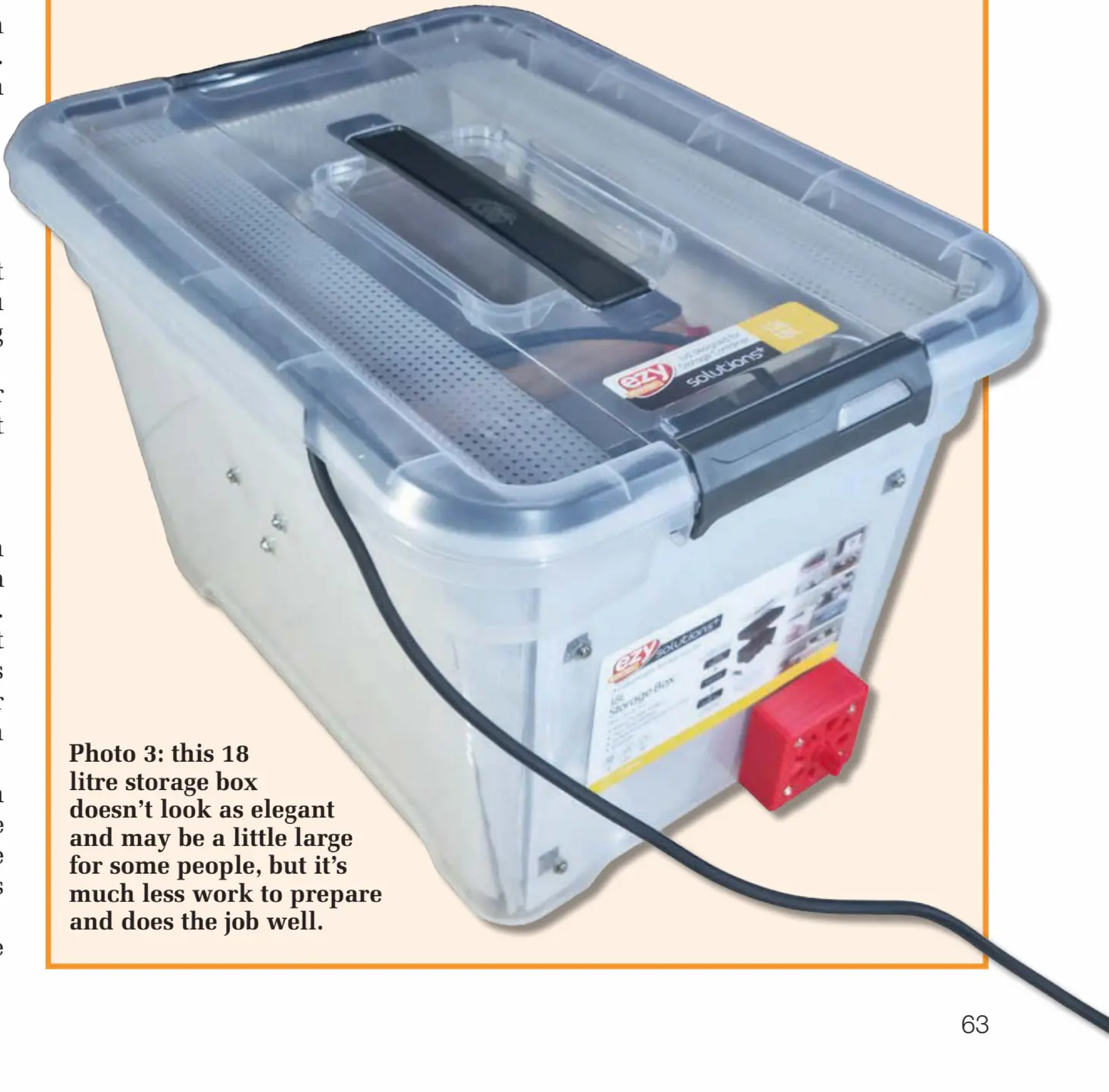


Photo 3: this 18 litre storage box doesn’t look as elegant and may be a little large for some people, but it’s much less work to prepare and does the job well.

installed inside the Filament Dryer, as that simplifies the wiring, and the temperature sensor is on the board. This means the controller will be operating at up to 50°C, perhaps a little more. That fine for most electronic components, but you will notice that we have specified high-temperature electrolytic capacitors and allowed for heatsinks on transistors Q1 and Q2.

Circuit details

The circuit is shown in Fig.2. An 8-bit PIC16F15214 operates as the timer, while an LM336-2.5 voltage reference (REF2) is used to produce a 2.5V reference, which is buffered by half of an LM358 op amp (IC1a). This is used in the temperature measurement circuit. The reason we have chosen the LM336-2.5 is it produces a reference voltage that is very stable over a wide temperature range.

The LM336-2.5 has a variation of just 6mV over 0-70°C, so we can expect to see an error of less than a degree in temperature control over our operational range.

The temperature sensor itself is a simple 1N4148 silicon diode (D6), using its -2.1mV/°C temperature coefficient. This is stable, reliable and used in many measurement circuits. The controller is a 'Bang-Bang' style, which simply turns the heating element on and off rather than implementing fancy control loops. This choice is again to keep things simple and cheap.

The controller comprises half of the LM358 (IC1b), which compares the voltage across the sense diode to the temperature set voltage. We use the 2.5V reference voltage to set the current through the sense diode via a 4.7kΩ resistor. The same reference

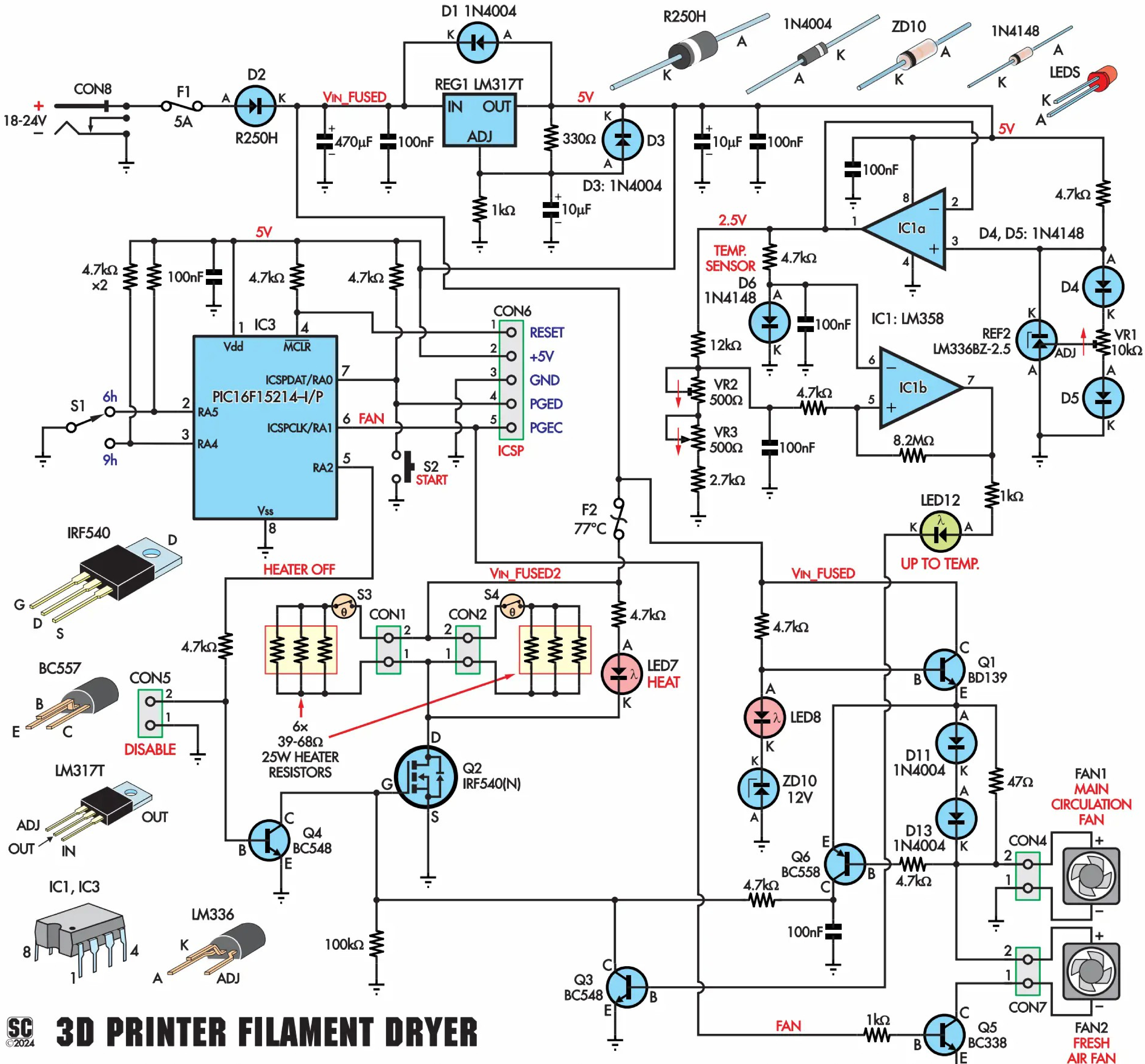

D1 1N4004
K
A
R250H
1N4004
ZD10
1N4148
LEDS
CON8
+
18-24V
−
F1
5A
D2
R250H
VIN_FUSED
470µF
100nF
REG1 LM317T
IN
OUT
ADJ
5V
330Ω
D3
D3: 1N4004
1kΩ
10µF
10µF
100nF
100nF
2.5V
IC1a
4.7kΩ
D4, D5: 1N4148
TEMP. SENSOR
D6
1N4148
4.7kΩ
100nF
IC1: LM358
REF2
LM336BZ-2.5
ADJ
D4
VR1
10kΩ
D5
12kΩ
VR2
500Ω
VR3
500Ω
2.7kΩ
4.7kΩ
100nF
IC1b
8.2MΩ
1kΩ
LED12
UP TO TEMP.
4.7kΩ ×2
100nF
4.7kΩ
4.7kΩ
IC3
Vdd
MCLR
PIC16F15214-I/P
ICSPDAT/RA0
ICSPCLK/RA1
RA5
RA4
RA2
Vss
6 FAN
S1
6h
9h
CON6
RESET
+5V
GND
PGED
PGEC
ICSP
S2
START
F2
77°C
IRF540
G
D
S
HEATER OFF
VIN_FUSED2
S3
S4
CON1
CON2
4.7kΩ
LED7
HEAT
BC557
B
E
C
4.7kΩ
CON5
DISABLE
6× 39-68Ω 25W HEATER RESISTORS
Q2
IRF540(N)
LM317T
ADJ
OUT
IN
OUT
Q4
BC548
IC1, IC3
LM336
K
A
ADJ
100kΩ
Q3
BC548
VIN_FUSED
4.7kΩ
Q1
BD139
LED8
ZD10
12V
D11
1N4004
D13
1N4004
47Ω
4.7kΩ
Q6
BC558
4.7kΩ
100nF
FAN1
MAIN CIRCULATION FAN
CON4
CON7
FAN2
FRESH AIR FAN
FAN
1kΩ
Q5
BC338
SC ©2024
3D PRINTER FILAMENT DRYER

Fig.2: the circuit of the Filament Dryer Controller. REF2 and IC1a create a 2.5V reference (trimmed by VR1). This biases diode D6, the temperature sensor. The voltage across D6 and the setpoint from VR2/VR3 are compared by op amp IC1b to drive Mosfet Q2 for powering the heating elements. Microcontroller IC3's timer limits the heating time and powers the fresh air fan periodically.

voltage generates the set voltage using trimpots VR1 and VR2 plus a couple of padder resistors.

By using this very stable 2.5V reference, we can be assured that the current through the sense diode and the set voltage are constant over time and temperature.

At room temperature, there is 400μA flowing through the sense diode, giving 0.56V across it. With the 12kΩ and 2.7kΩ padders and two 500Ω potentiometers, we get a temperature set point range of about 20-50°C. The reason we have included two pots is to allow us to use one (VR2) to set the minimum temperature to room temperature, while the other (VR3) is used to choose the temperature setpoint.

With trimpot VR2 at the nominal value of 220Ω, the minimum voltage will be 0.489V (2.5V × 2.90kΩ ÷ [12kΩ + 2.92kΩ]). The maximum voltage will be 0.554V (2.5 × 3.42kΩ ÷ [18kΩ + 3.42kΩ]). The difference is 0.065V, and at 2.1mV/°C, that gives a spread of 31°C.

Even using 1% resistors, the errors in the voltage divider are significant. If one is 1% high and the other is 1% low, the setpoint could move as much as 7°C. By adjusting VR2 so the minimum setpoint is room temperature, we can calibrate such errors out.

The output of IC1b is low when the sensed temperature is below the setpoint and goes high when the temperature exceeds the setpoint. The 8.2MΩ resistor adds about 2°C of hysteresis by feeding back the output voltage to slightly shift the setpoint voltage.

The ratio of the 8.2MΩ and 4.7kΩ resistors results in a shift of just a couple of milivolts, which is what we need. This stops IC1b from oscillating once the setpoint is reached.

With the controller being flat out on or off, and the degree or two of hysteresis, the temperature control is not super precise. But for warming the filament to dry it out, that is OK.

For the timer, we started by considering simple CMOS timer circuits and the venerable 555. To get a nine-hour period from these is not easy, so the cheapest way to make the timer was to use a PIC. These cost nearly $1.50 in single units, a fraction of the cost of the discrete solution, and can be programmed to do a huge range of jobs.

We consider the timer to be an integral part of this design and strongly recommend against omitting it.

Parts List – Filament Drying Chamber

1 double-sided PCB coded 28110241, 126 × 93mm
1 18-24V DC 3A+ power supply (eg, laptop charger)
2 12V DC 40mm fans, 10mm-thick **[Farnell 3794004]**
1 40mm fan grille **[Farnell 055025]**
2 PCB-mounting M205 fuse clips (for F1)
1 5A 250V M205 fuse (F1)
1 77°C axial thermal fuse (F2) **[Farnell 3996011]**
5 2-pin vertical polarised headers, 2.54mm pitch (CON1-2, CON4-5, CON7)
5 2-pin polarised header plugs with pins
1 5-pin header, 2.54mm pitch (CON6; **optional**, for programming IC3 in-circuit)
1 PCB-mounting DC socket, 2.1mm ID or to suit power supply plug (CON8)
1 PCB-mounting 90° miniature SPDT toggle switch (S1) **[Farnell 9473300]**
1 PCB-mounting 90° sub-miniature SPST pushbutton switch (S2) **[Farnell 2008742]**
1 10kΩ side-adjust single-turn trimpot (VR1)
1 500Ω side-adjust single-turn trimpot (VR2)
1 500Ω 16mm single-gang linear potentiometer (VR3)
2 TO-220 micro-U heatsinks (**optional**) **[Farnell 1611445]**
2 90°C normally-closed (NC) thermal switches/breakers (S3, S4) **[Farnell 1006844]**

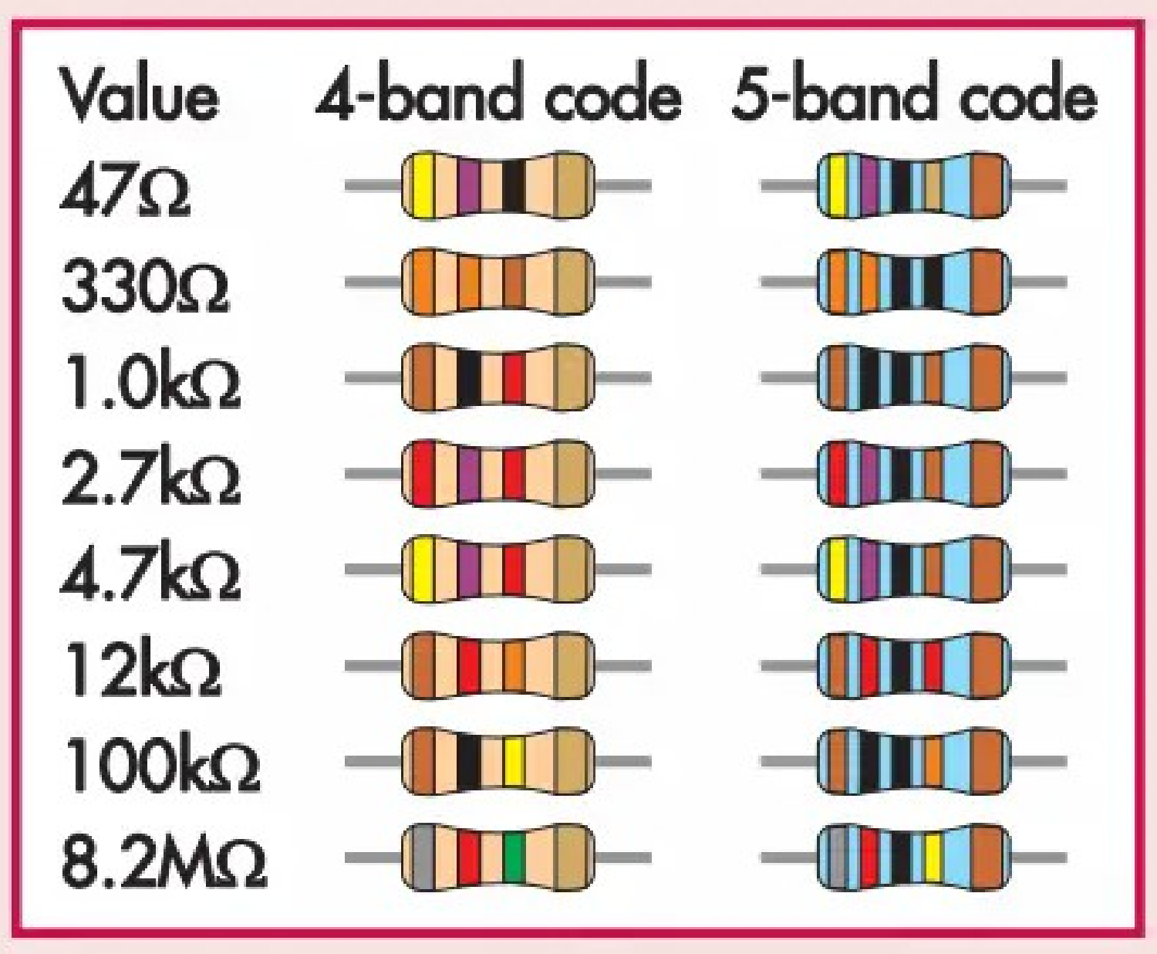
Hardware (common to both versions)
1 3D-printed vent ("Vent Rotor.STL", "Vent Rotor Base.STL" & "Vent No Fan.STL")
1 3D-printed fan cover ("Fan Shroud.STL")
6 M3 × 25mm panhead machine screws
18 M3 hex nuts & 32 M3 flat washers
1 3m length of high-temperature (90°C+) heavy-duty hookup wire
1 250mm length of 6mm diameter heatshrink tubing
1 2m length of 5-10mm wide open-cell foam adhesive tape
1 small tube of thermal paste

Hardware (for plastic box version)
1 large polypropylene box
2 1.5mm-thick aluminium plates, 210 × 180mm
Panhead machine screws: 8 M3 × 6mm, 32 M3 × 10mm, 8 M3 × 16mm, 6 M3 × 25mm
Tapped spacers: 4 M3 × 15mm, 16 M3 × 25mm male/female hex spacers
Other: 58 M3 shakeproof washers, 46 M3 hex nuts

Hardware (for timber box version)
2 3D-printed handles ("Filament Dryer Rail Tall.STL")
1 sheet of 12mm MDF or plywood
1 1.5mm-thick aluminium plate, 330 × 225mm
Panhead machine screws: 6 M3 × 6mm (30 if building lid), 16 M3 × 10mm, 4 M3 × 16mm, 24 M3 × 25mm, 1 M4 × 10mm (for attaching handle to lid)
Tapped spacers: 12 M3 × 6mm (for lid), 10 M3 × 15mm
Other: 42 M3 shakeproof washers, 38 M3 hex nuts

Capacitors
1 470μF 35V 105°C electrolytic
2 10μF 50V 105°C electrolytic
7 100nF 50V multi-layer ceramic or MKT

Semiconductors
1 LM358 dual single-supply op amp, DIP-8 (IC1)
1 LM336BZ-2.5 voltage reference diode, TO-92 (REF2)
1 PIC16F15214-I/P 8-bit microcontroller programmed with 2811024A.HEX, DIP-8 (IC3)
1 LM317T adjustable positive linear regulator, TO-220 (REG1)
1 BD139 80V 1.5A NPN transistor, TO-126 (Q1)
1 IRF540(N) 100V 30A N-channel Mosfet or similar, TO-220 (Q2)
2 BC548 30V 100mA NPN transistors, TO-92 (Q3, Q4)
1 BC338 25V 800mA NPN transistor, TO-92 (Q5)
1 BC558 30V 100mA PNP transistor, TO-92 (Q6)
4 1N4004 400V 1A diodes (D1, D3, D11, D13)
1 R250H or 6A10 400V 6A diode (D2)
3 1N4148 75V 200mA diodes (D4-D6)
1 12V 0.4W or 1W zener diode (ZD10)
2 5mm red LEDs (LED7, LED8)
1 5mm green LED (LED12)



Value	4-band code	5-band code
47Ω		
330Ω		
1.0kΩ		
2.7kΩ		
4.7kΩ		
12kΩ		
100kΩ		
8.2MΩ		

Resistors *(all ¼W 1% axial unless noted)*
1 8.2MΩ   1 100kΩ   1 12kΩ   12 4.7kΩ
1 2.7kΩ   3 1kΩ   1 330Ω   1 47Ω
6 39Ω (18V), 47Ω (19-20V) or 68Ω (24V) 25W aluminium body resistors **[Ohmite HS25 series]**

Our dryer includes two fans. The first is to circulate air inside the box and it runs full-time. There is also a ventilation fan that runs briefly every 10 minutes. This is intended to draw fresh air into the box and to exhaust the hot (and possibly moist) air. This ventilation fan is driven by the PIC microcontroller.

We do not want to continuously change the air in the enclosure, as it would require a lot of power to keep the temperature elevated. So our tiny PIC microcontroller drives the vent fan sparingly.

Software

The program in the timer is quite simple. At power-up, the PIC goes into an idle state, disabling the heater and ventilation. It stays in this state until the user presses the start button. This requires a deliberate action by the user.

Once the start button is pressed, the timer moves into the running state. If IC3's RA4 digital input is low, the timer drives its RA2 output low and counts nine hours. If RA5 is low instead, the output is low for six hours. After the selected time, the heater is switched off and the system goes back to the idle state. If the input is invalid, it remains idle.

The PIC includes a secondary timer that drives digital output RA1 to switch on the ventilation fan every 10 minutes.

The timer output and the output of the temperature sensor comparator are combined using open-collector transistors Q3 and Q4, which disable heater drive transistor Q2 when they are on. When the box is up to temperature, the output of IC1b goes high, switching on Q3, which disables the heater. Green LED12 is in series with this output, and lights showing that the set temperature has been achieved.

Switching the load on is implemented using an IRF540 or similar power Mosfet with a gate pullup resistor to 12V. The gate drive pullup is derived from the ventilation fan power supply, which might seem an odd choice. The ventilation fan draws current through D11, D13 and the parallel 47Ω resistor.

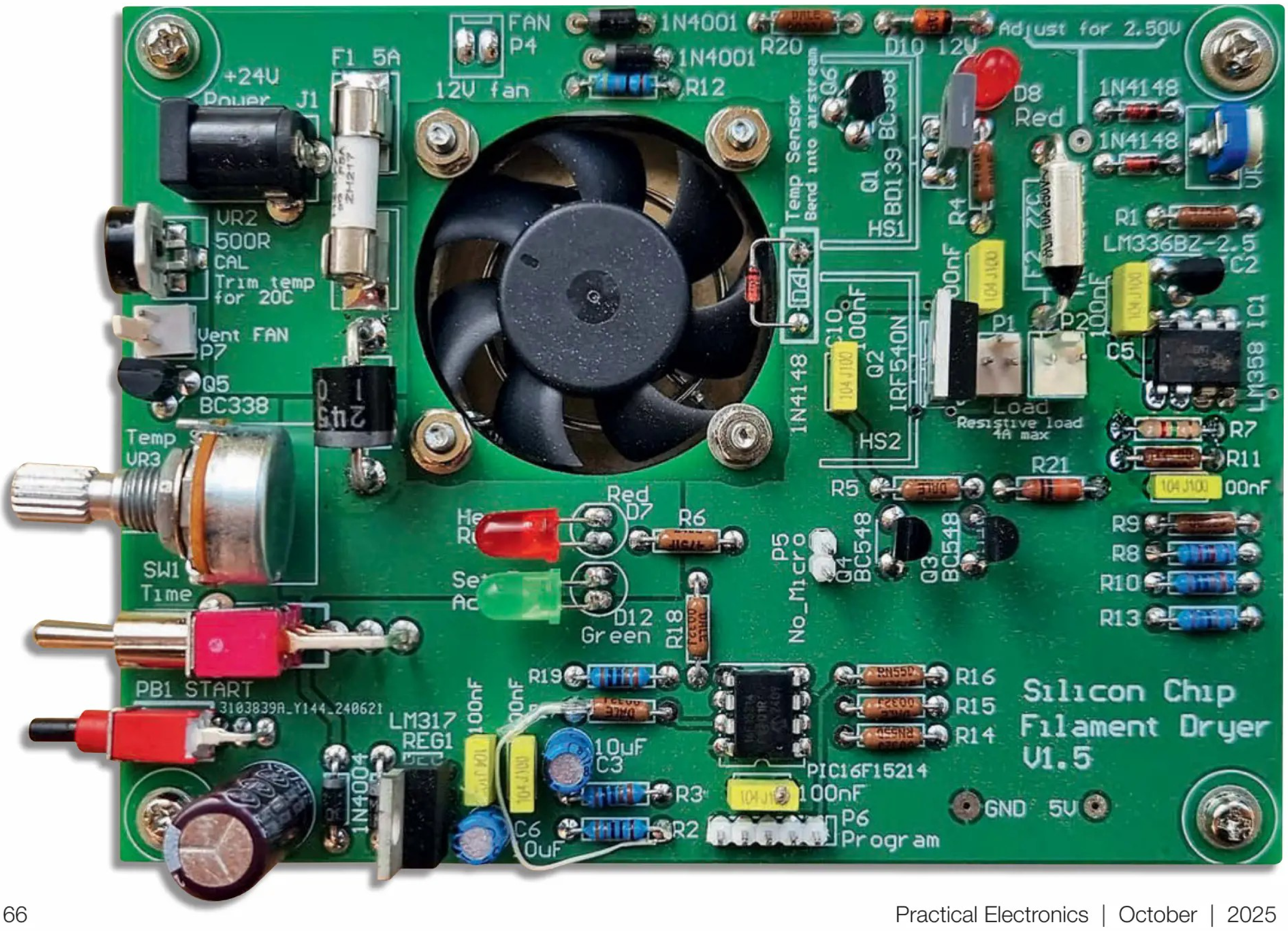
The specified fan draws 60mA in operation and develops 1.2V across these diodes. This voltage switches on Q6 on via its 4.7kΩ base resistor, which forms the Mosfet gate drive.

If the fan stalls, its internal controller reduces its supply current to 2mA and attempts to restart it every few seconds. This 2mA current only generates 94mV across the 47Ω resistor, which is not enough to switch Q6 on, and consequently the Mosfet gate drive is removed. Thus, we disable the heater if the ventilation fan is stalled or not working.

For Q2, pretty much any TO-220 package, low-$R_{DS(ON)}$ N-channel Mosfet will work. They virtually all have the same pinout. If you want to use a different Mosfet from our recommended part, look for one with an $R_{DS(ON)}$ under 0.1Ω.

For example, the MTP3055V has an $R_{DS(ON)}$ of 0.18Ω and for a load current of 3A, it will dissipate 1.6W (3A^2 × 0.18Ω). That would demand the use of a flag heatsink; there is room for this on the PCB. The recommended IRF540 has an $R_{DS(ON)}$ of 0.077Ω and will dissipate 0.7W at 3A (or 0.4W for the IRF540N version), which will make it warm but it won't require a heatsink.

Photo 4: the top side of the prototype PCB. The fan is mounted to the underside using four M3 x 16mm machine screws with matching hex nuts.


FAN
P4
12V fan
1N4001
1N4001
R20
R12
D10 12V
Adjust for 2.50V
+24V
Power
J1
F1 5A
Temp Sensor
Bend into airstream
Q6
BC558
Q1
BD139
HS1
R4
D8
Red
1N4148
1N4148
VR2
500R
CAL
Trim temp
for 20C
Vent FAN
P7
Q5
BC338
R1
LM336BZ-2.5
C2
C10
100nF
Q2
IRF540N
1N4148
P1
P2
C5
IC1
LM358
Load
Resistive load
4A max
HS2
R7
R11
R21
R5
100nF
Temp
VR3
SW1
Time
Red
D7
R6
P5
No_Micro
Q4
BC548
Q3
BC548
R9
R8
R10
R13
D12
Green
R18
PB1 START
3103839A_Y144_240621
LM317
REG1
1N4004
100nF
R19
10uF
C3
R3
R2
C6
PIC16F15214
100nF
P6
Program
R16
R15
R14
Silicon Chip
Filament Dryer
V1.5
GND
5V

There are two headers for wiring up the heater resistors. This allows you to run separate wiring to two banks of resistors, making the wiring and layout easier in some builds. The current rating of the recommended headers is 3A, so you could get away with using just one.

We have included a thermal fuse in the power supply to the Mosfet. The specified fuse has a current rating of 10A AC but in our application, we are breaking nominally 2A DC. The fuse does not have a DC rating but that is well within its capacity. This device will fuse at 77°C, and will hold at 55°C continuously. Should your enclosure exceed 55°C for extended periods, you may trigger this protection.

The heater

We performed a number of tests on the boxes we're presenting and determined that we need 50W to heat our enclosures to 50°C reliably in a 20°C room. This is also a good maximum, as per the safety considerations we touched on earlier.

To allow us to spread the power around the enclosure, we are using six 25W resistors mounted to large heatsinks. We have used 68Ω devices, which at 24V will dissipate 50W in total. To spread the heat, we used a 330 × 225mm aluminium sheet folded to fit inside our timber box, or two 210 × 180mm panels for the plastic box.

If using a 19V supply, the heating resistor values need to be reduced to 47Ω to keep that 50W target.

We recommend the cheapest aluminium case power resistor we could find, mentioned in the parts list. The cost is around $20 for six, so you can save some decent money reusing parts you have. It is important that the devices you select can be bolted to the heat spreader, as this ensures they do not get hot enough to create a hazard.

We tried using 10W ceramic resistors each dissipating 5W. While they were operating within their specification, their surface temperature of over 130°C would have the potential to create a hazard if combustible material fell onto them.

Next month

The second and final article next month will have the construction and testing details, including building or adapting and then insulating the container. **PE**

Safety considerations for the Filament Dryer

In designing the controller, we undertook a hazard assessment and developed controls for each hazard we identified, seeking to mitigate these hazards as much as reasonably practicable. This in broader engineering often forms part of a "Safety Engineering Program". This process involves identifying credible hazards them applying the 'hierarchy of controls' which, in order, are:

- Eliminate the hazard
- Substitute to avoid / minimise the hazard
- Apply engineering controls
- Add administrative controls (how it is used)
- Use Personal Protective Equipment (PPE)

In safety engineering, there is an important differentiation between a hazard, which is a potential outcome, and the risk this represents, which considers the likelihood of this occurring. The intention of applying the hierarchy of controls is to mitigate and minimise the overall risk of a system.

Our hazard assessment was undertaken to inform the design of the project and to shape the solution, both to minimise the underlying hazards in the design and also to apply substitutions, engineering and administrative controls to further mitigate residual risks. By keeping a record of the approaches to managing safety, and building those into the design, we can then test the project to ensure that these controls do what we expect.

The hazards and controls we identified for the filament dryer are shown in Table 1 (Hazard & Risk Assessment). Some significant changes in design were implemented. Those practised in the safety art will note that we have picked parts of a larger process to document here, as a full safety program is comprehensive and at times less than fascinating. We have, however, included some important elements for your consideration when making your own version of this.

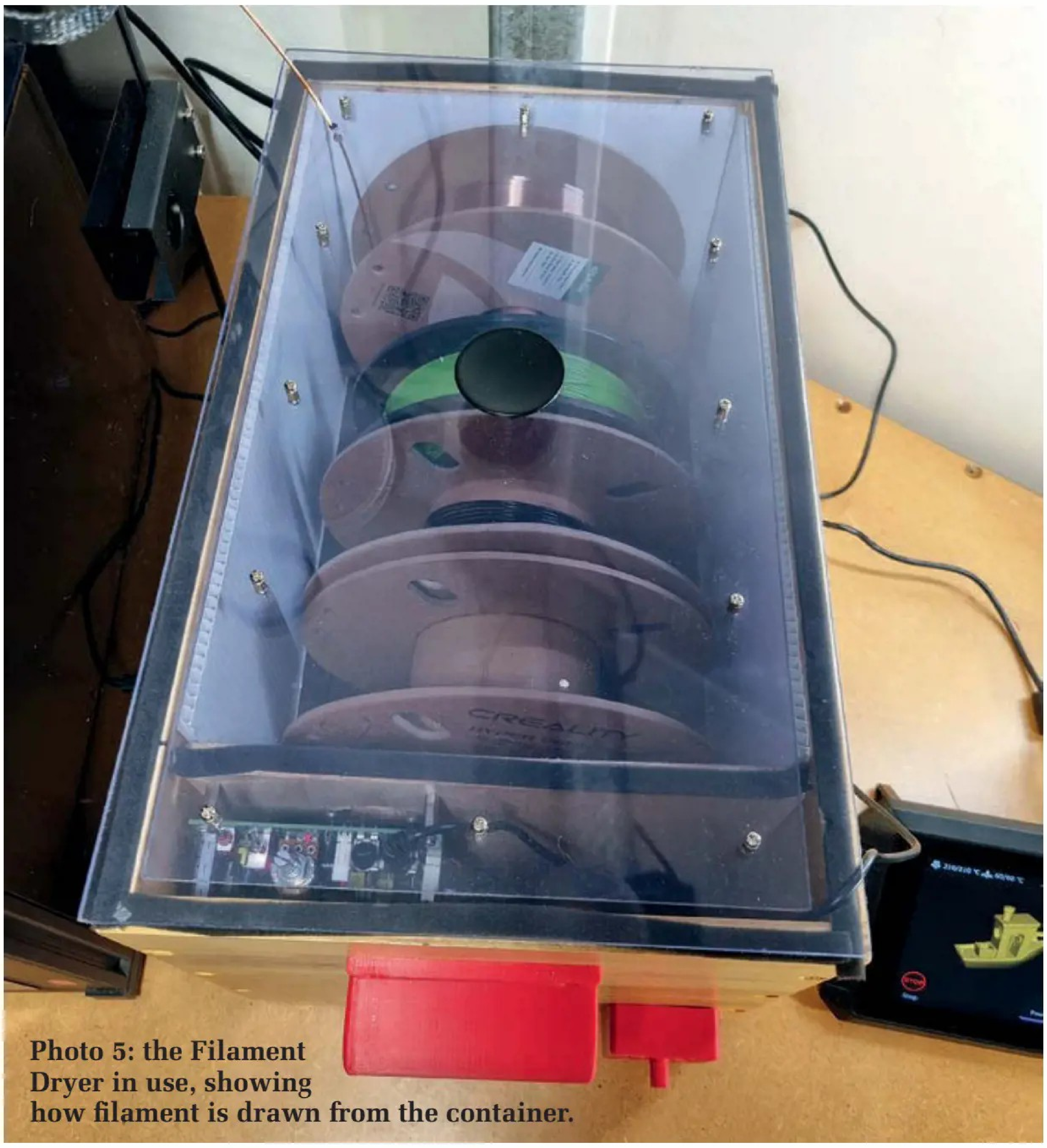
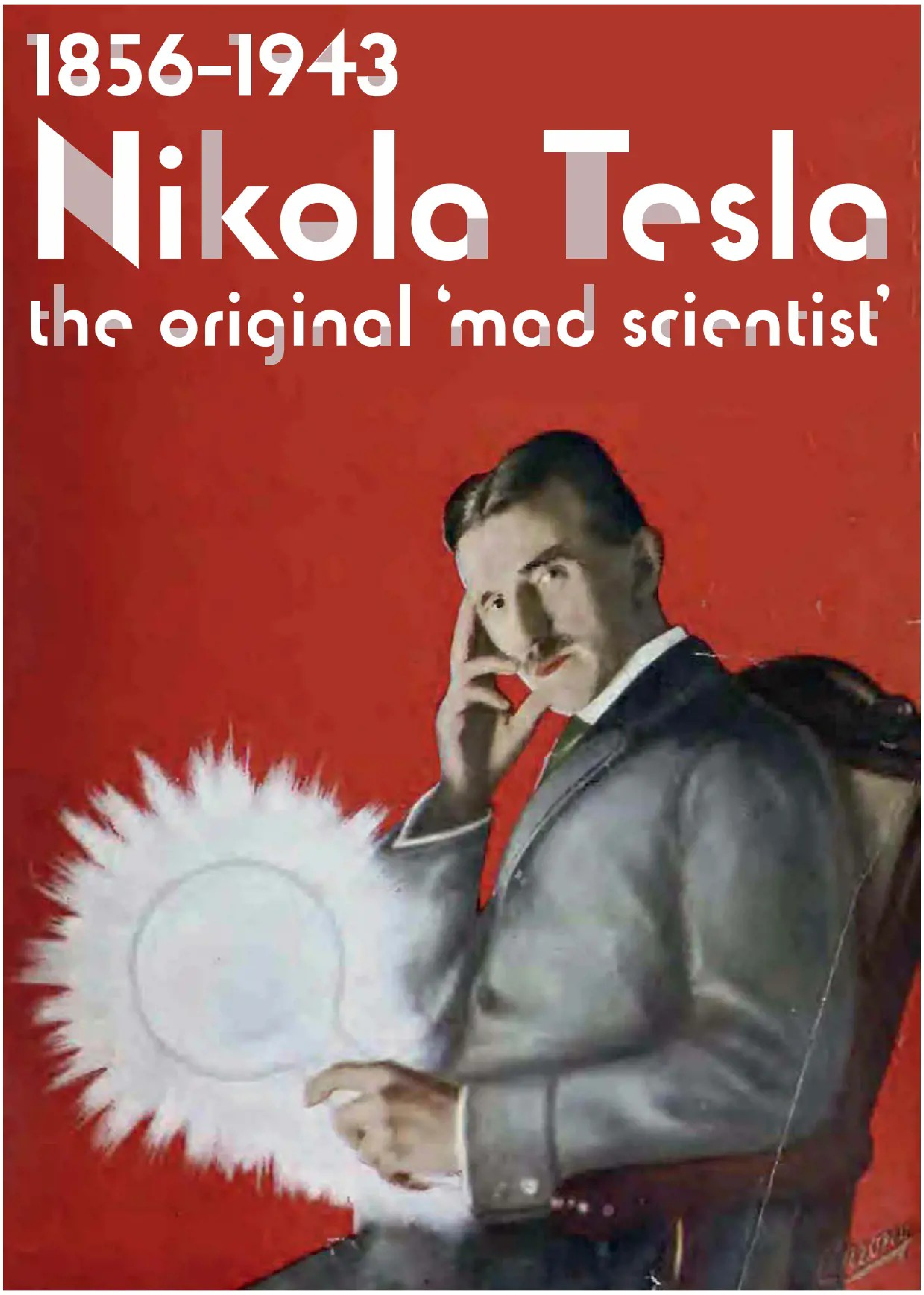


Photo 5: the Filament Dryer in use, showing how filament is drawn from the container.

1856–1943

Nikola Tesla

the original 'mad scientist'



Nikola Tesla was a prolific inventor, engineer, futurist and essayist. He spoke eight languages, had a wide range of interests and has been described as a "Renaissance man". Despite his 'mad scientist' vibe, his contributions to our modern industrial civilisation are significant.

Part 1 by Dr David Maddison, VK3DSM

Tesla on the cover of Electrical Inventor magazine, February 1919. The lead image is based on a photo of Tesla from around 1900 demonstrating wireless power transmission. He is holding a partially evacuated glass bulb that's glowing due to the electric field from a nearby Tesla coil. See https://w.wiki/AZMz

Befitting someone who made such contributions, he was said to have been born during a violent lightning storm at midnight between July 9th and 10th, 1856, in Croatia. According to his family, the midwife said the lightning was a "bad omen" and that he would be a "child of darkness", to which the mother replied, "No. He will be a child of light."

Fortunately, he became a force for good and lived for 87 years. He passed away in 1943, leaving a remarkable and world-changing legacy, which we will now examine.

Of his numerous inventions & developments, among the most important were his contributions to three-phase AC electricity, the induction motor, the Tesla coil (used in many early radios) and one of the world's first hydroelectric power plants. The development of three-phase electricity allowed the transmission of electrical power over long distances, one basis of modern industrial civilisation.

The electric car company Tesla is named after him. Two museums are dedicated to him, and there are statues of him on Goat Island, USA and Queen Victoria Park, Canada, both near Niagara Falls. There are also several Tesla memorial plaques in Manhattan, New York, USA.

Tesla's thought processes

Tesla's creative genius might be attributable to his unusual thought processes. These facilitated his ability to visualise and create things. He wrote in Electrical Experimenter, February 1919:

> *In my boyhood I suffered from a peculiar affliction due to the appearance of images, often accompanied by strong flashes of light, which marred the sight of real objects and interfered with my thought and action. They were pictures of things and scenes which I had really seen, never of those I imagined... I was quite unable to distinguish whether what I saw was tangible or not.*
>
> *Then I observed to my delight that I could visualize with the greatest facility. I needed no models, drawings or experiments. I could picture them all as real in my mind. Thus I have been led unconsciously to evolve what I consider a new method of materializing inventive concepts and ideas, which is radically opposite to the purely experimental and... so much more expeditious and efficient.*

The Life of Nikola Tesla, part one

... My method is different. I do not rush into actual work. When I get an idea I start at once building it up in my imagination. I change the construction, make improvements and operate the device in my mind. It is absolutely immaterial to me whether I run my turbine in thought or test it in my shop. I even note if it is out of balance. There is no difference whatever, the results are the same.

In this way I am able to rapidly develop and perfect a conception without touching anything. When I have gone so far as to embody in the invention every possible improvement I can think of and see no fault anywhere, I put into concrete form this final product of my brain. Invariably my device works as I conceived that it should, and the experiment comes out exactly as I planned it.

Also, like many creative geniuses, he had various eccentricities.

Tesla, mathematics and quantitative theories

Tesla was different from most other scientists and engineers. His writings are highly descriptive and contain few equations. He was a visual thinker and performed mathematics 'visually' rather than presenting it through formal methods.

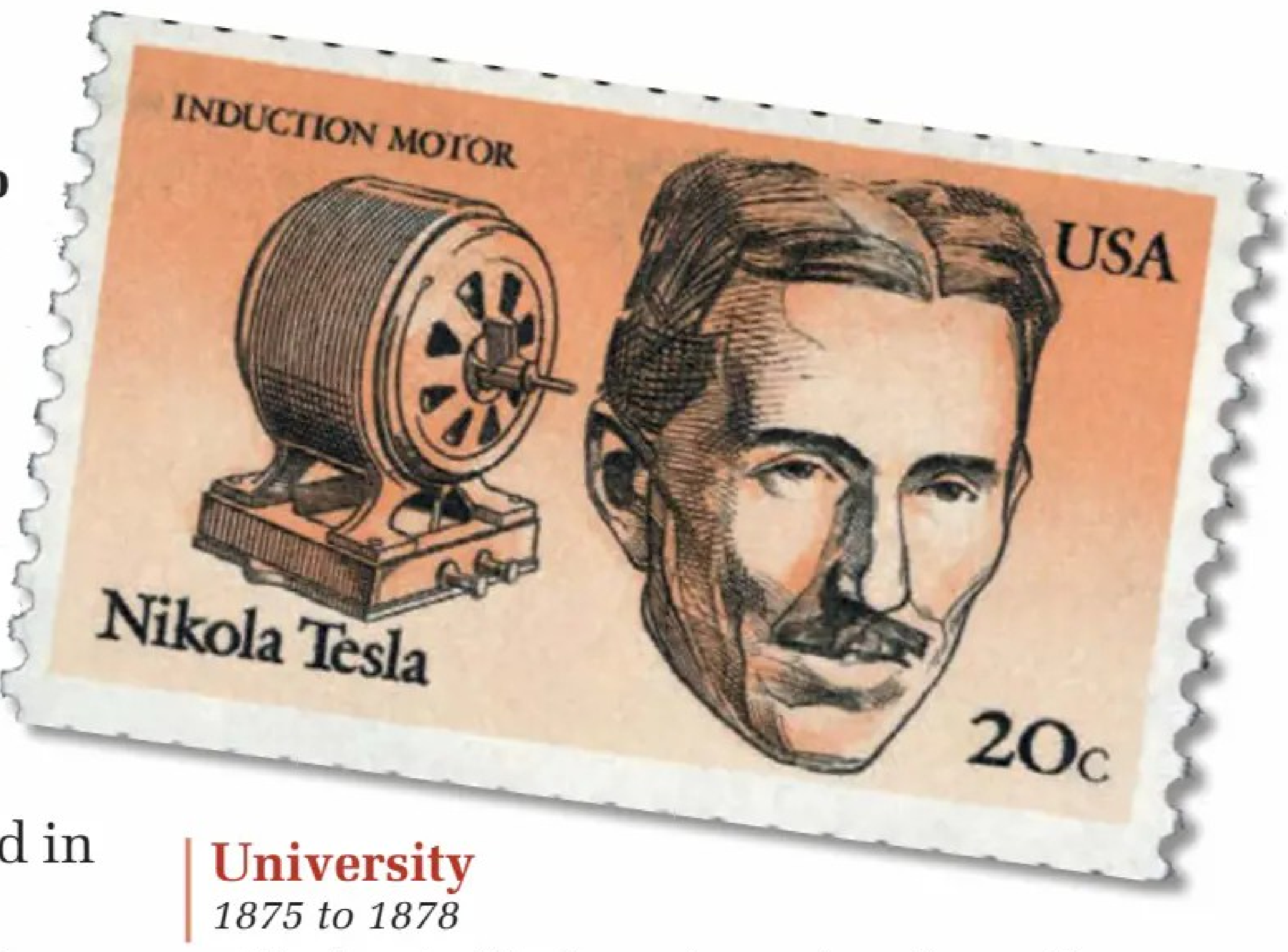
He also did not accept Maxwell's equations, which are the basis of electricity, magnetism and optics. ("We can no longer believe in the Maxwellian hypothesis of transversal ether-undulations and the literal truth of its corollaries." – https://pemag.au/link/aby0).

He generally ignored quantitative theories; it has been suggested he may have suffered from "mathematical aphasia". His work was experimental, usually practical, descriptive and not analytical. In that respect, he was much like Michael Faraday; it was Maxwell who turned Faraday's intuitive ideas into equations.

Separating fact from fiction

While Tesla's contributions to technology were undoubtedly outstanding, it should be recognised that work has been attributed to Tesla that either he did not originate or where he was a partial contributor.

Also, during his lifetime, there was fierce commercial competition regarding which electrical supply technology was to be adopted, so there were often claims and counterclaims made that didn't necessarily reflect the reality of who invented what. Tesla had no marketing department; he had to do his own promotion, which is often reflected in his writing style.



An American postage stamp featuring Nikola Tesla. *Source:* https://postalmuseum.si.edu/object/npm_2008.2007.74

Not all of Tesla's inventions or ideas were successful or viable. Tesla's early work, such as with the induction motor, generators and the Tesla coil was excellent. However, some of his later work, which involved long-distance wireless electricity transmission, was not based on sound physical principles.

Tesla's patents

Tesla was prolific and obtained around 112 US patents, 29 UK patents and six Canadian patents. He applied for 33 patents that were not granted. He also had patents in other countries for a total of around 300; for a complete list, see https://w.wiki/AZLY

Tesla's life and career

We will now take a look at some of Tesla's milestones in chronological order. This article will end in 1897; the remainder will be covered in the second and final article in this series, to be published next month.

University
1875 to 1878

Tesla studied engineering from September 1875 at the Graz University of Technology in Austria but, having started with excellent results, did not finish his degree. He left after the first semester of the third year, apparently losing interest in his studies while spending too much time in a café and associated activities.

He sat no exams that year and was excluded. While at university, he saw a Gramme dynamo, which operated either as a generator or motor. He conceived a way to eliminate the commutator, which his professor didn't believe was possible.

This ultimately led to Tesla's development of the AC induction motor, which contains no commutators. He received an honorary doctorate from Graz in 1937.

Prague
1880

Tesla arrived in Prague and spent much of his time reading at the Klementinum Library and Národní Kavárna

Interesting facts about Nikola Tesla

- He had a great sense of humour.
- He was a rival of Edison, not a sworn enemy; they had a mutual respect for each other.
- He had the idea of a 'smartphone' type device in 1901. He described to his then-backer J.P. Morgan a handheld device he said would deliver stock quotes and telegram messages.
- For unknown reasons, he hated pearls and would not speak to any lady wearing them.
- He had a photographic memory.
- He had a fear of germs, always wore white gloves and rarely shook hands.
- He asked for large numbers of napkins at meals.
- He never stayed in a room or floor number divisible by three.
- He ran his life according to a strict daily schedule.
- He was very particular about dress and grooming.
- He had a beloved pet pigeon.
- The SI unit for magnetic flux is named after him, the Tesla (T).
- Toward the end of his career, he ran out of profitable ideas, or at least people who were prepared to back him financially. As a result, he passed away in poverty with many unpaid debts.

café. He also attended lectures at the University of Prague but was not enrolled as a student.

Budapest Telephone Exchange

1881

In 1881, Tesla commenced work with the Budapest Telephone Exchange, a new company that was not yet functional. So he helped set it up, as a draftsman and later chief electrician, making several design improvements.

Continental Edison Company

1882

In 1882, Tesla worked for Edison in Europe. He started by installing lighting systems, but his expertise was noted, and he became involved in designing improved dynamos and motors.

Rotating magnetic fields

1882

The idea of a rotating magnetic field was conceived as early as 1824 by François Arago but, according to Tesla, he conceived of its use in an AC electric motor while walking through a park in Budapest in 1882 (documented on p198 of the PDF at https://pemag.au/link/aby0).

Although he doesn't explicitly mention the rotating magnetic field, it was the basis of the motor. His idea of eliminating commutators and the rotating magnetic field combined to create the AC induction motor.

Prototype induction motor

1883

In 1883, while working for Edison in Strasbourg, he constructed (on his own time) an induction motor but could not find any interest in it.

Emigration to the USA

1884

Tesla's manager in Europe was recalled back to Edison in the USA and requested Tesla to come to work at the Edison Machine Works in New York City. There, he managed staff involved in installing New York's electricity utilities. He was also involved in developing an arc lamp street lighting system, but that needed high voltages and was incompatible with the Edison system.

Tesla's designs were not utilised; there had been improvements in incandescent lighting. Tesla only worked there for six months before he left, apparently after a dispute about an alleged promised bonus.

Tesla Electric Light & Manufacturing

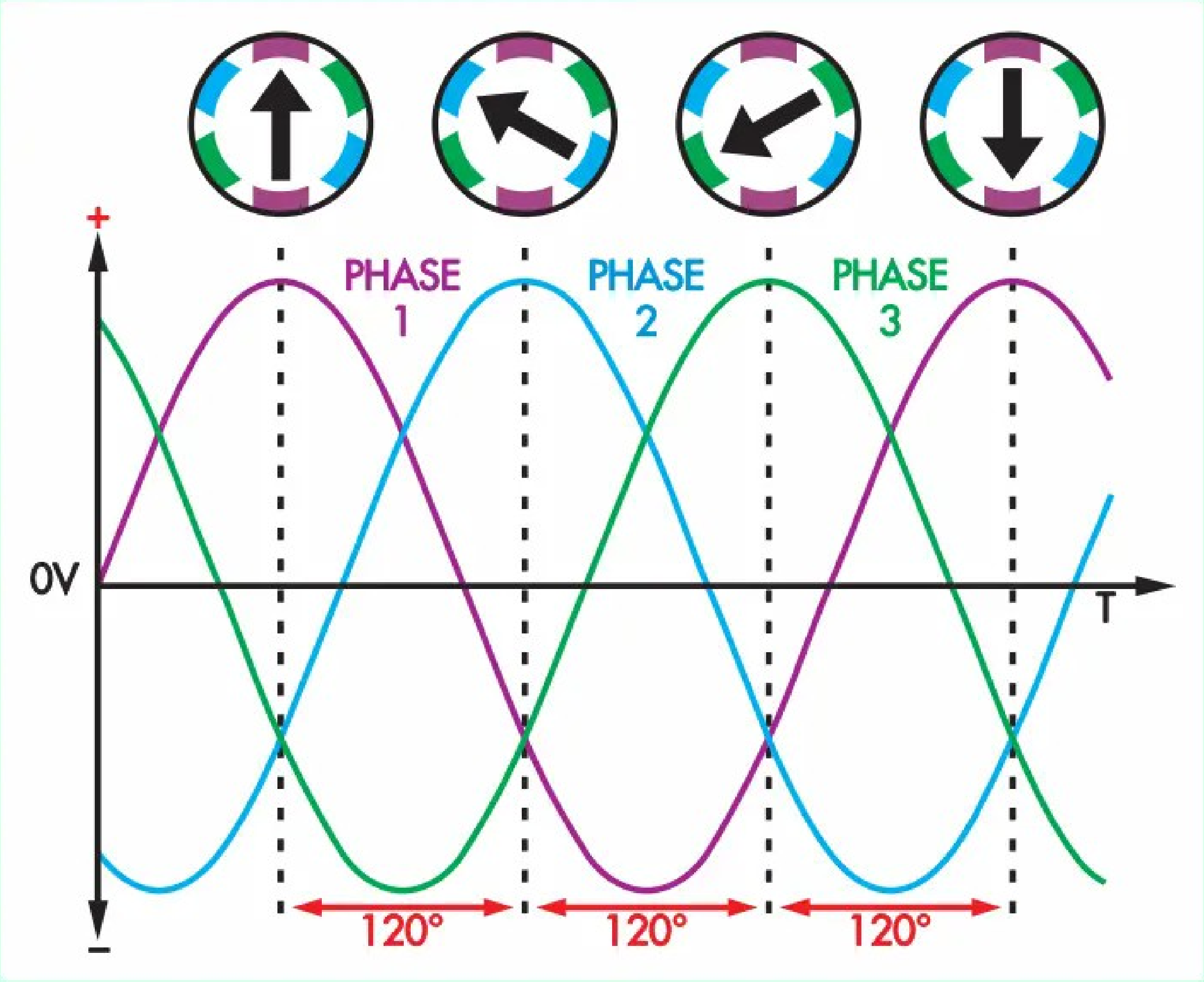
1885

After leaving Edison, investors asked Tesla to design a system of electric arc lamps for lighting the streets of New York and other cities. This led to the establishment of the Tesla Electric Light and Manufacturing Company.

What is polyphase electricity?

Many early writings on AC electricity use the term "polyphase". Polyphase refers to an AC electrical system with two or more AC voltage supplies supplied by separate wires and with the sinewaves of each displaced from each other by a certain amount, usually described in degrees. Early work on polyphase systems was with two phases, but today, three phases is the most common configuration.

A three-phase system is twice as efficient at conductor utilisation as a single-phase system. Polyphase power, especially three-phase, is ideal for induction motors, as it can easily generate a rotating magnetic field, eliminating high-maintenance commutators and allowing simple and inexpensive construction.



The principle of a rotating magnetic field in a three-phase induction motor. The magnetic field sequentially rotates between the various motor poles, causing the rotor to follow it and rotate.

Tesla continued obtaining patents for motors, generators and other equipment, but the investors showed no interest in those. They decided manufacturing was too competitive and just wanted to run an electric utility.

They left the company, which left Tesla penniless; worse, he had assigned his patents to them in return for the now-worthless stock. He described it as "the hardest blow" he ever received.

Digging ditches

1886 to 1887

After the failure of his company, he made a living digging ditches.

Labs in New York

1887 to 1902

During this period, Tesla maintained a series of laboratories in Manhattan, New York. They were on Liberty Street (1887-1889), Grand Street (1889-92), South Fifth Avenue (1892-95) and East Houston Street(1895-1902).

The Tesla Electric Company

1887

In 1887, with new investors, Tesla set up the Tesla Electric Company and the Liberty Street laboratory. In the same year, he invented an induction motor (patented in 1888) that would run on the newly developed AC system. It was becoming popular in Europe because of its advantages of long-distance transmission with little electrical loss.

The motor used polyphase current which, at the time, was two-phase (we have three now). The polyphase current generated a rotating magnetic field. The advantage of this motor was that it did not need a commutator, which caused sparks, required high maintenance, and was expensive and complex. Apart from motors, Tesla developed generators and other power system devices.

Polyphase induction motor patent

1888

In 1888, Tesla obtained US patent 381,968, the first of a series on electric motors (it continued until 1896). It was for commutator-free polyphase alternating current induction motors (see Fig.1). He envisaged two- and three-phase motors in that patent.

He also published descriptions of other motors, including a synchronous motor for which the rotation speed is locked to the AC power frequency. Those are ideal for clocks and other motors where precise speed control is essential.

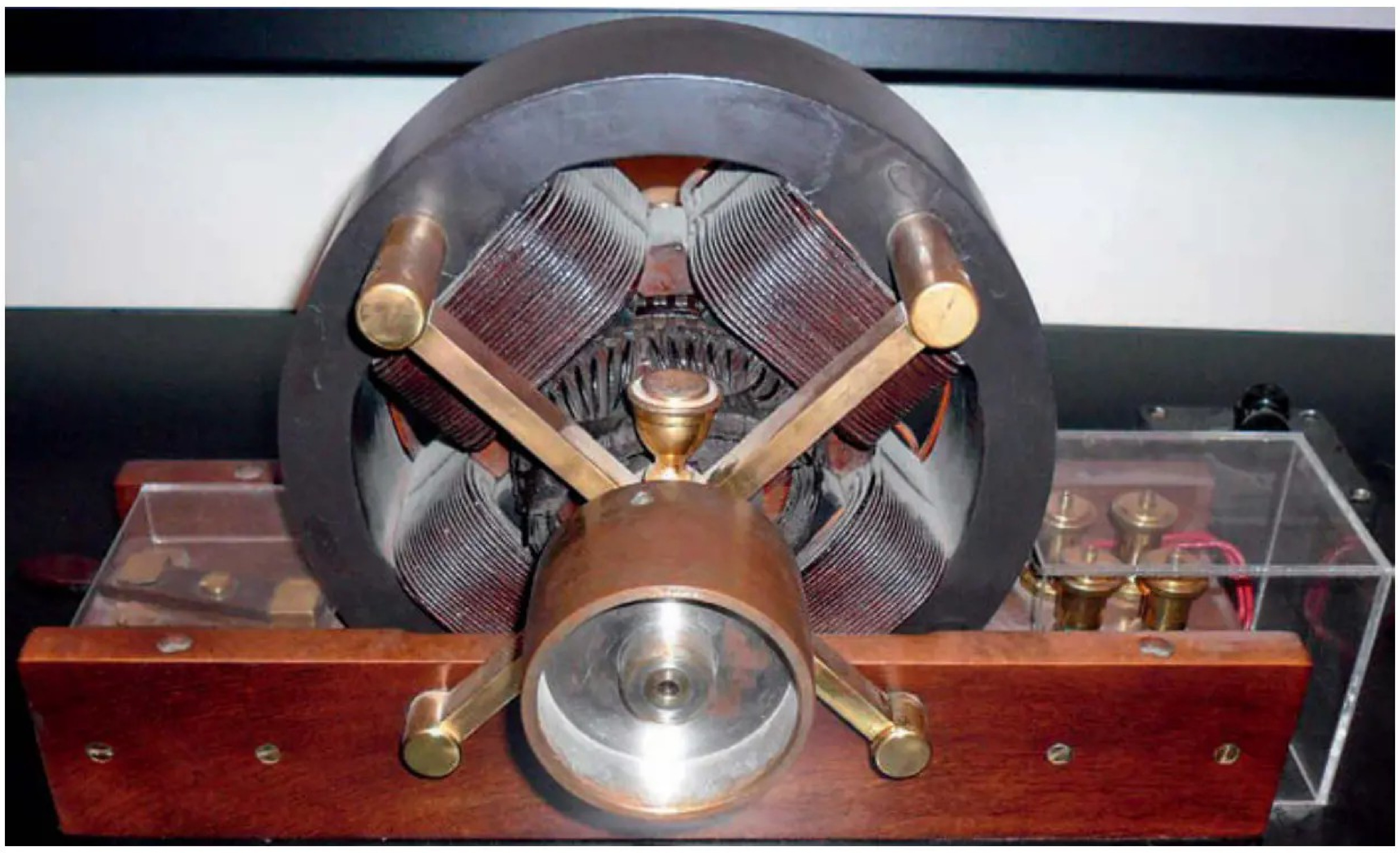


Fig.1: a model of Tesla's first induction motor at the Tesla Museum, Belgrade, Serbia. *Source:* https://w.wiki/AZM$

Galileo Ferraris independently invented and demonstrated a commutator-free two-phase alternating current induction motor in 1885, but he didn't patent it because he could see no practical application.

Westinghouse
1888

George Westinghouse of the Westinghouse Electric & Manufacturing Company was already marketing an AC power system and needed a suitable AC motor. He considered using Ferraris' motor but decided that Tesla's was superior.

Tesla's investors negotiated with Westinghouse in 1888 to license his AC transformer, dynamo and motor designs for cash and stock plus a royalty per horsepower of AC motor sold. He also hired Tesla as a consultant for a hefty fee.

During 1888, there was intense competition between the three main electrical companies: Westinghouse, Edison and the Thomson-Houston Electric Company. There was also the emerging "war of the currents" between the AC system promoted by Westinghouse and the DC system promoted by Edison.

Tesla's motor was not immediately successful, and the adoption of the polyphase AC system was limited. The intense competition meant that Westinghouse did not have the resources to continue to develop Tesla's induction motor or the polyphase AC system.

Westinghouse was then in serious financial trouble. He explained the difficulties to Tesla, and in 1891, Tesla released Westinghouse from the royalty clause on the motors and he later purchased the patent. The cancellation of the royalty clause meant that Tesla would get a minute amount of the true value of his motor and generator patents.

Polyphase current and generators
1888

The first of two important patents this year was US patent 390,413 for a "System of Electrical Distribution" for electrical transmission of polyphase power such that "two or more circuits may have a single return path or wire in common".

The second was US patent 390,414 on a "Dynamo Electric machine" concerning adapting existing dynamos easily and cheaply to polyphase alternating current.

A large number of patents
1888 to 1891

This period was enormously productive for Tesla; many patents were granted, including 43 US patents in the area of single and polyphase currents, induction and other types of electric motors and generators.

Independently wealthy
1889

Tesla became independently wealthy due to Westinghouse licensing his patents, so he had the funds from 1889 to pursue his own interests. It has been suggested that Tesla was not a particularly good businessman and was always looking for investors, unlike Edison. Also, Tesla tended to work for himself, while Edison employed many other people and had multiple projects on the go at once.

Wireless lighting
1890

In 1890, Tesla started experimenting with wireless lighting and performed public demonstrations with power transmitted by inductive or capacitive coupling. This work continued for about another ten years.

Tesla coil
1891

In 1891, Tesla patented a type of resonant transformer that is now known as the Tesla coil (US patent 454,622). A resonant transformer uses capacitors across one or more windings, which act as coupled resonant tuned circuits. It produces high-voltage, low-current, pulsed or AC electricity at radio frequencies. Voltages produced can range from 50kV to millions of volts at 50kHz to 1MHz.

The essential elements of a Tesla coil are an air-cored 'oscillation transformer', a capacitor, a high voltage primary transformer and a spark gap.

Tesla used these coils in numerous experiments and built them to very large sizes, such as in Colorado and Wardenclyffe. Experiments Tesla used the coils for included investigating biological effects, high-frequency phenomena, lighting (for which the

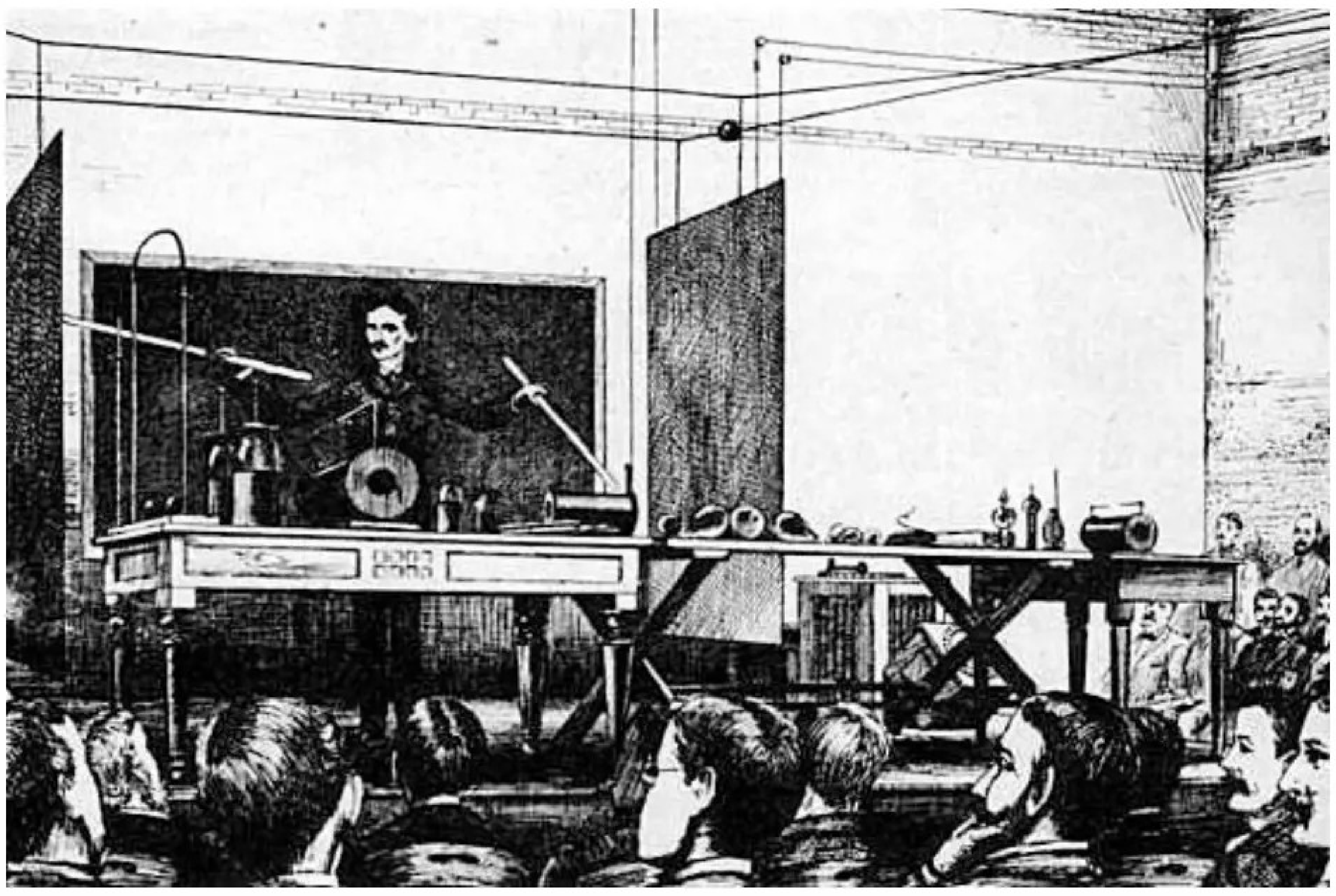


Fig.2: Tesla giving a demonstration of wireless power transmission in 1891. *Source:* https://w.wiki/AZN2

original patent was issued), phosphorescence, radio, wireless power transmission and X-rays.

Tesla made a radio antenna out of the high-voltage end of the secondary part of the transformer, turning it into a radio transmitter. Such an arrangement was used in most early spark-gap radios for wireless telegraphy applications until the 1920s, when the vacuum tube rendered them obsolete.

Lighting power supply
1891

In 1891, he applied for and was granted US Patent 454,622 for a means of generating high-voltage and high-frequency electricity for lighting purposes.

Incandescents & power transmission
1891

In this year, he obtained US Patent 455,069 for an incandescent light. On May 20th, Tesla demonstrated wireless power transmission to the American Institute of Engineers in a lecture hall at Columbia University. The lecture was entitled "Experiments with alternate currents of very high frequency and their application to methods of artificial illumination".

In one demonstration, he vertically suspended two large zinc sheets from the ceiling, which were connected to a high-frequency, high-voltage Tesla coil. He held an unconnected gas-filled tube between them, and the tube glowed due to the electrostatic field between the sheets, just as a fluorescent tube glows when near a high-voltage power line due to capacitive coupling.

Wireless power transmission
1891 to 1898

Tesla's dream was global wireless electrical transmission. From 1891 to 1898, he performed numerous experiments and demonstrations in wireless transmission via capacitive or inductive coupling (see Fig.2). In 1899, he commenced larger-scale experiments at Colorado Springs and later Wardenclyffe.

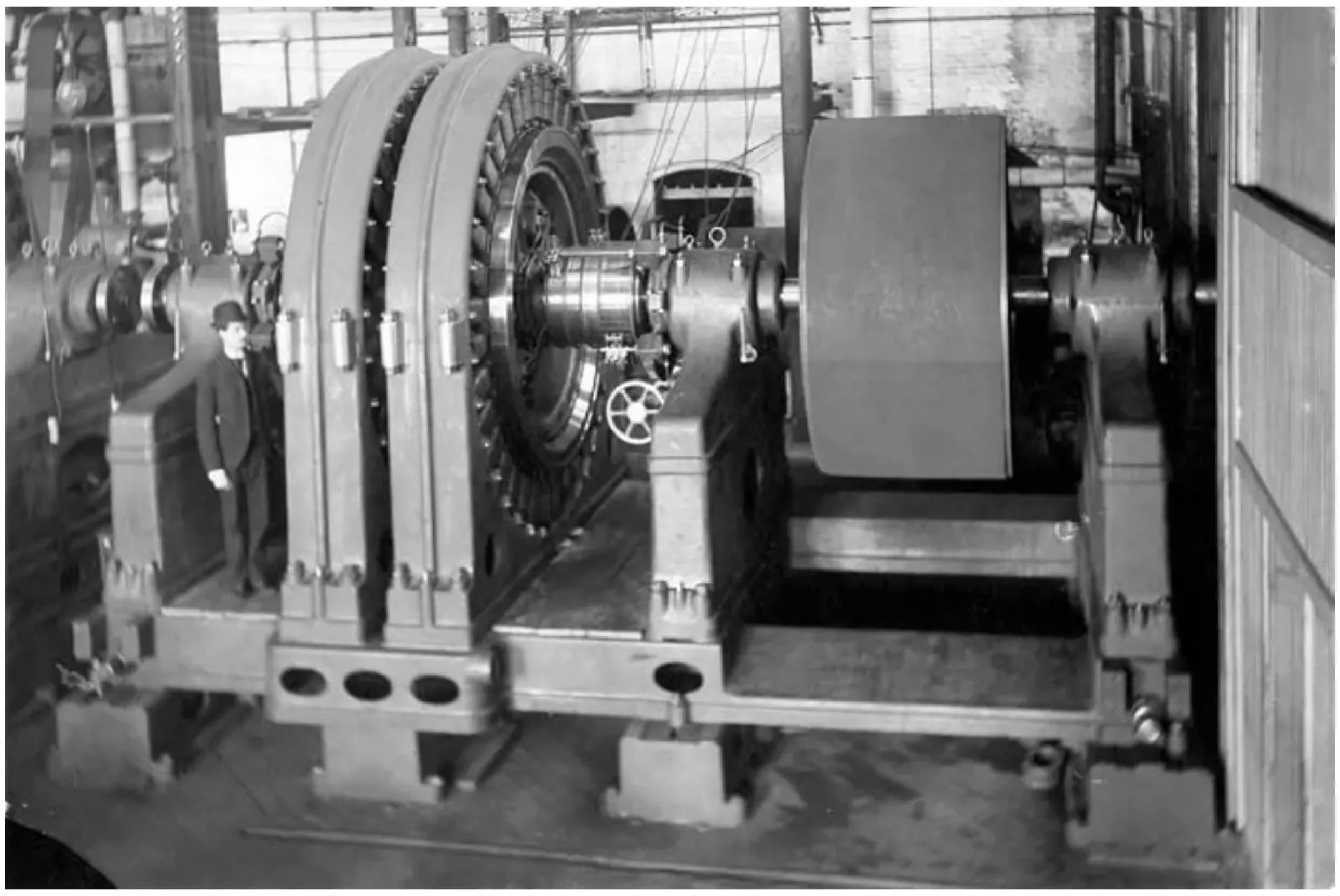


Fig.4: in Tesla's design, two single-phase alternators were magnetically coupled, 90° out-of-phase to provide two-phase AC for the exposition lighting. Note the alternator's size in relation to the man. *Source:* https://historicpittsburgh.org/islandora/object/pitt:20170320-hpichswp-0011

AIEE organisation
1892 to 1894

From 1892 to 1894, he was vice president of the American Institute of Electrical Engineers, a forerunner of the IEEE.

Visit to Europe
1892

He gave a series of lectures in London and Paris on "Experiments with alternate currents of high potential and high frequency".

Chicago World's Fair
1893

Also called the World's Columbian Exposition, was a significant turning point in the "war of the currents", with George Westinghouse winning the lighting contract ($399,000) over Edison's DC system ($554,000) – see Fig.3.

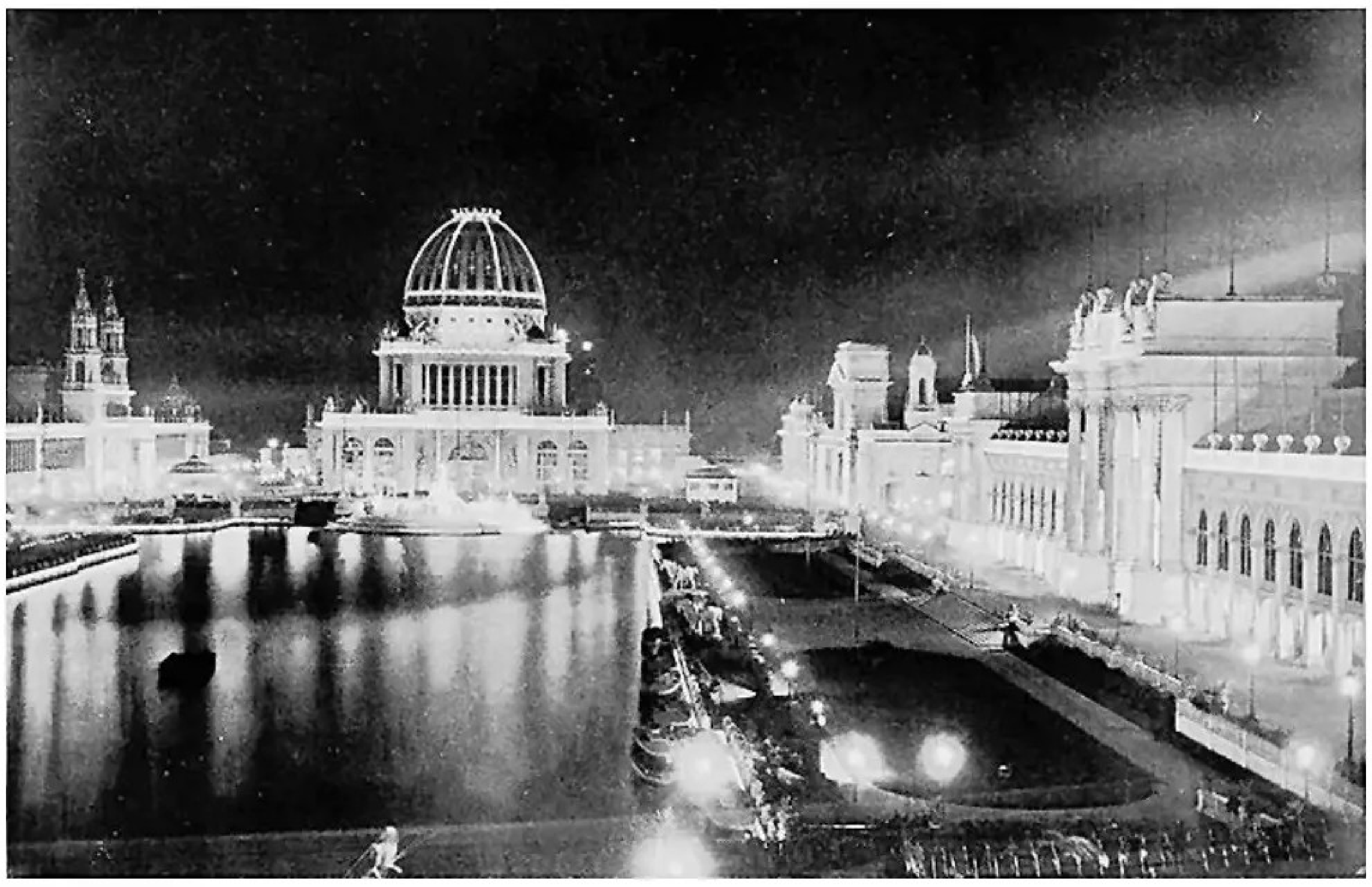


Fig.3: night-time lighting at the 1893 Chicago World's Fair using Tesla's patented AC and lighting systems. *Source:* https://w.wiki/AZN3

Westinghouse used Tesla's AC power patents to power lighting of their own design (they could not use Edison's lights). The lighting and other systems at the fair used twelve 745kW 60Hz single-phase AC generators of Tesla's design. These were mounted in pairs and arranged to provide two-phase power (see Fig.4).

The Westinghouse Company also had a section showcasing Tesla's inventions, such as induction motors (Fig.5) and generators. The rotating magnetic field used in induction motors was demonstrated with the "Egg of Columbus" (Fig.7).

Tesla demonstrated wireless lighting using neon tubes, although he did not invent neon lighting (see Fig.6). He also demonstrated clocks synchronised to the mains frequency.

Talks at Franklin Institute & NELA
1893

His talk was "On light and other high frequency phenomena" and he mentioned the "transmission of intelligible signals and power to any distance without the use of wires" (radio). He also discussed the idea of transferring power over long distances through the Earth.

Niagara Falls hydroelectric power
1893

In 1893, Tesla was invited to consult for the Niagara Falls hydroelectric

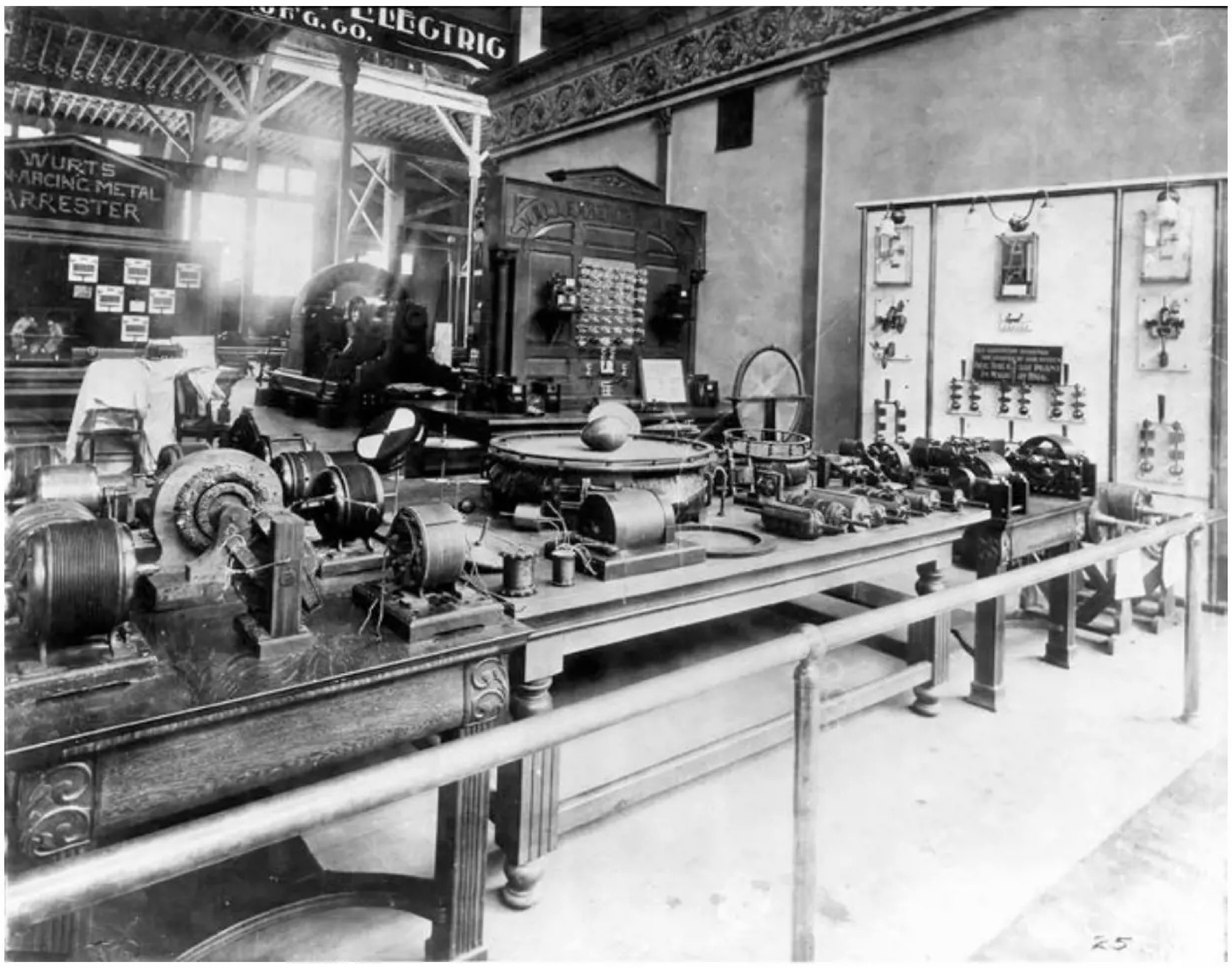

F'G. CO. ELECTRIC
WURTS
ARCING METAL
ARRESTER

Fig.5: an exhibition of Tesla's motors and the "Egg of Columbus" at the 1893 Chicago World's Fair. ***Source:*** **https://w.wiki/AZN4**

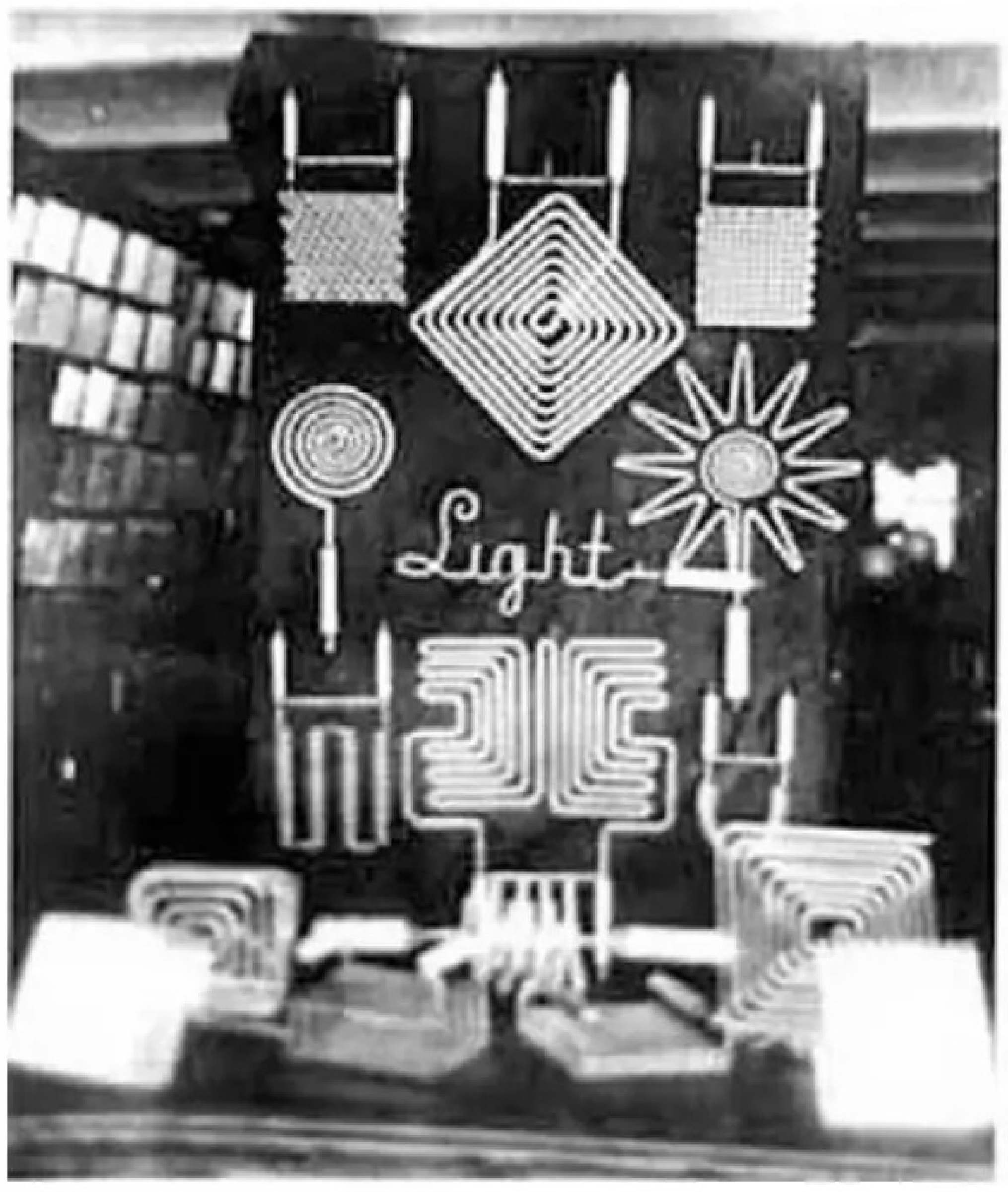


Fig.6: Tesla's wireless lighting demonstration using neon tubes at the Chicago World's Fair.

project. Proposals that had been put forward for the electrical system included two- and three-phase AC and high-voltage DC.

Tesla advised that a two-phase AC system from Westinghouse, designed by Tesla and based on his patents, was the best and most reliable option (see Fig.9). Westinghouse was awarded the main contract based on Tesla's advice and the success of the Tesla and Westinghouse displays and lighting system at the Columbian Exposition.

Nine of the twelve patents used for the plant's machinery were Tesla's. Electricity from the plant first went to a nearby factory in 1895 and then to Buffalo, New York, in 1896.

At a talk about the City of Buffalo receiving power from Niagara on January 12th, 1897, at the Ellicott Club, Tesla said:

> *It is a monument worthy of our scientific age, a true monument of enlightenment and of peace. It signifies the subjugation of natural forces to the service of man, the discontinuance of barbarous methods, the relieving of millions from want and suffering.*
>
> *From "The Age of Electricity" by Nikola Tesla, Cassiers Magazine – London, March 1897, pp378-386.*

This AC power plant is regarded as the final victory of the "war of the currents", with Tesla's AC proving itself superior to Edison's DC.

A low frequency of 25Hz was chosen, as it was expected that much of the power would be converted to DC via rotary converters for uses such as aluminium production. However, it was realised that three-phase power was superior for transmission efficiency, so phase-changing transformers were used to convert the two-phase power to three-phase.

These are known as "Scott-T" transformers since they were invented by Charles F. Scott, who worked for Westinghouse in the late 1890s. The configuration of this type of transformer is shown in Fig.12. The first output phase (0°) is a direct transformer-coupled copy of the first input phase (0°) via transformer T1.

The second phase at 120° is generated by connecting the centre tap of

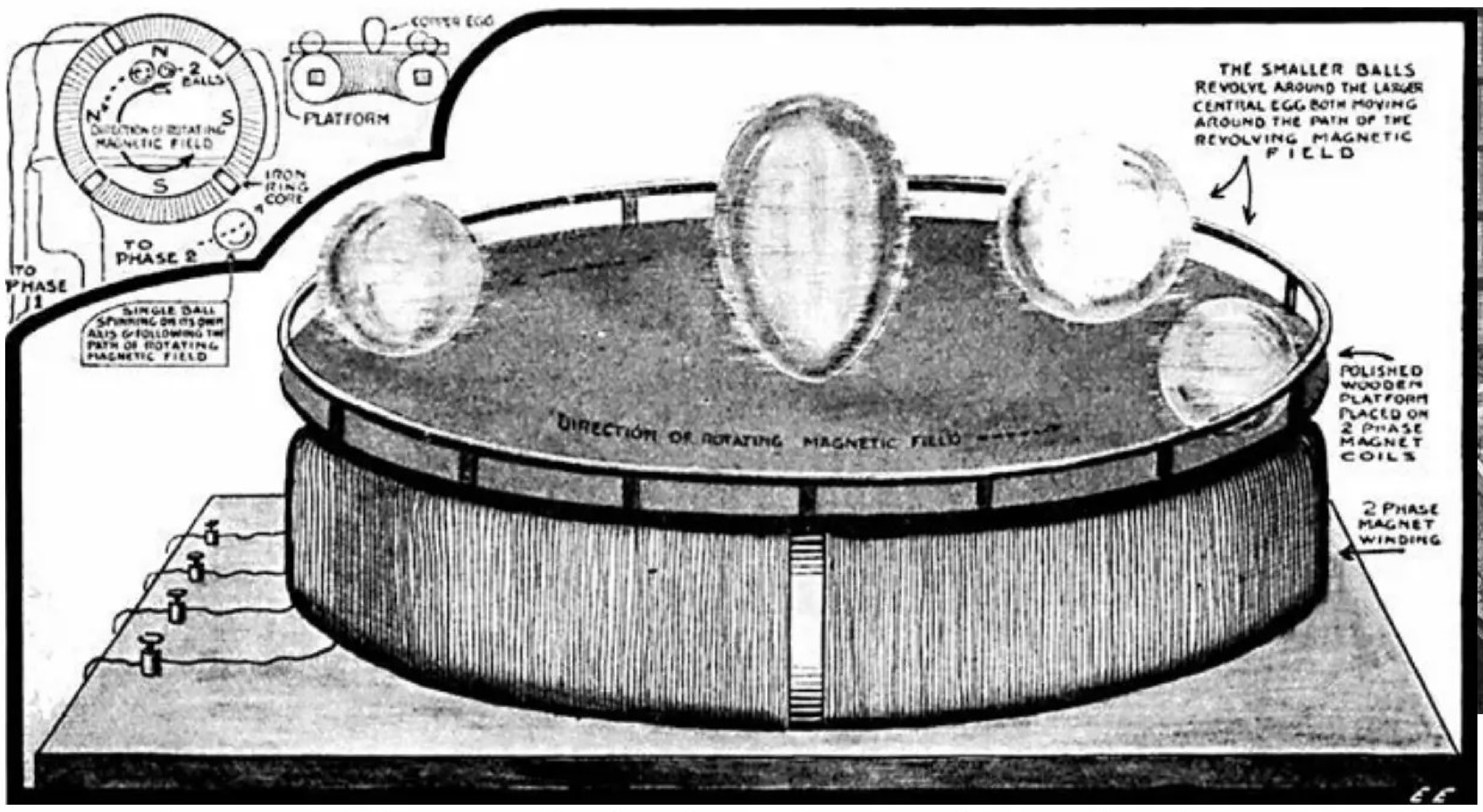

COPPER EGG
N
2 BALLS
N
S
DIRECTION OF ROTATING MAGNETIC FIELD
S
PLATFORM
IRON RING CORE
TO PHASE 2
TO PHASE 1
SINGLE BALL SPINNING ON ITS OWN AXIS & FOLLOWING THE PATH OF ROTATING MAGNETIC FIELD
THE SMALLER BALLS REVOLVE AROUND THE LARGER CENTRAL EGG BOTH MOVING AROUND THE PATH OF THE REVOLVING MAGNETIC FIELD
DIRECTION OF ROTATING MAGNETIC FIELD
POLISHED WOODEN PLATFORM PLACED ON 2 PHASE MAGNET COILS
2 PHASE MAGNET WINDING

Fig.7: a drawing of the "Egg of Columbus" that was designed to demonstrate the rotating magnetic field devised by Tesla. ***Source:*** **https://w.wiki/AZN5**

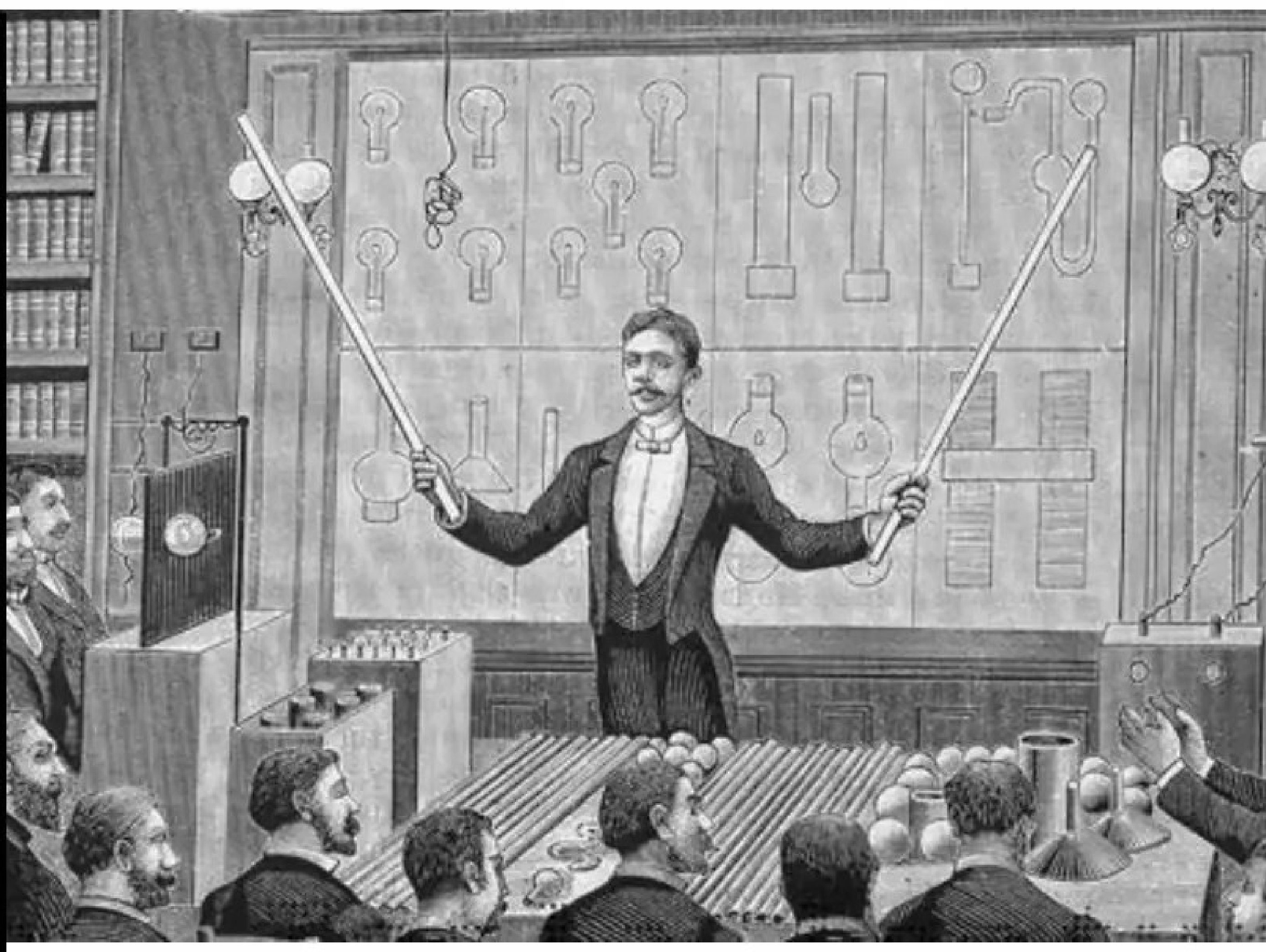


Fig.8: a drawing of Tesla lecturing before the French Physical Society and The International Society of Electricians in the 1880s.

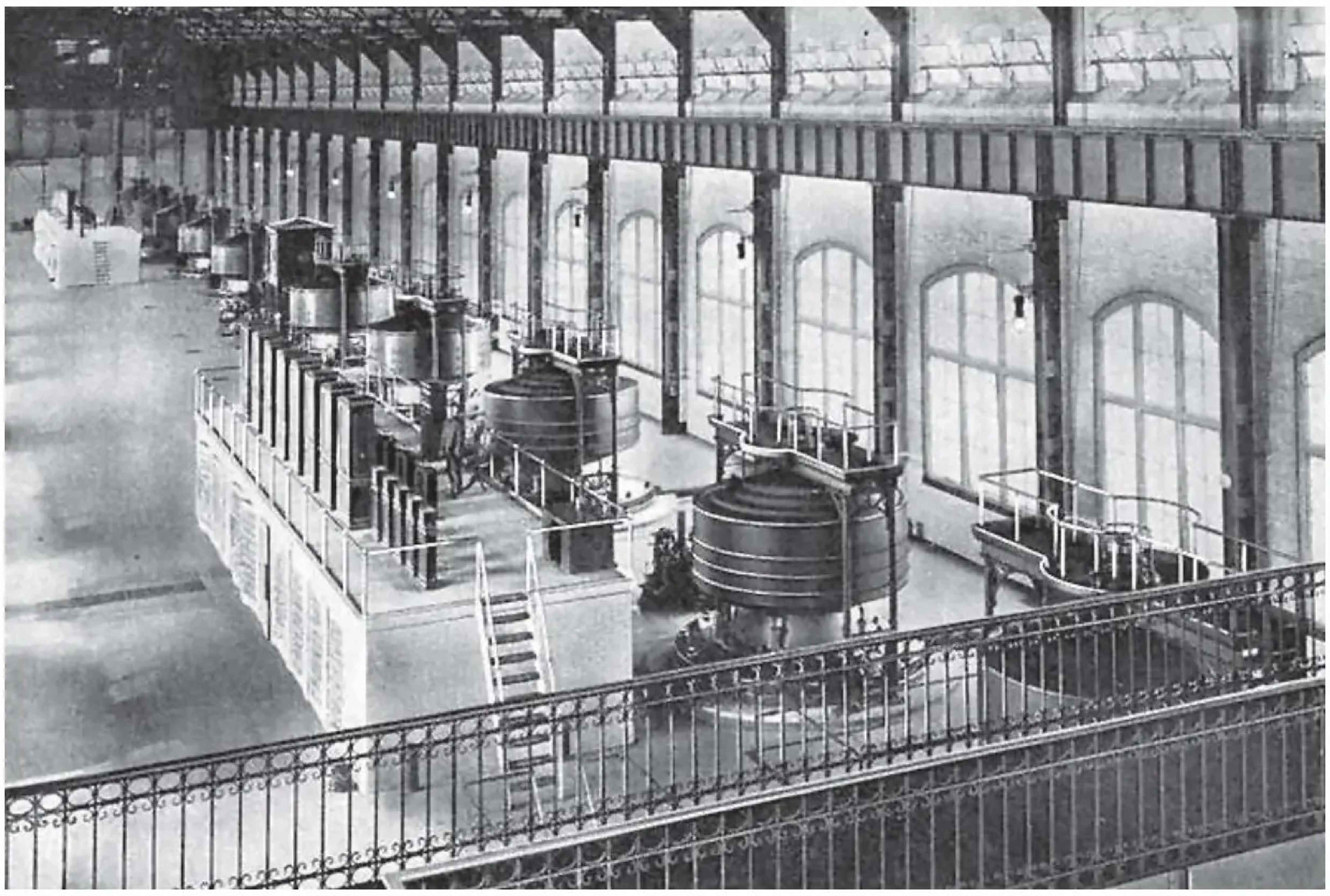


Fig.9: ten 3.7MW 25Hz 2kV Westinghouse generators at Edward Dean Adams Power Plant in Niagara Falls, installed in 1895. The voltage was stepped up to 10kV or 20kV depending upon how far away the destination was. These generators remained in use until 1961. *Source:* https://w.wiki/AZN6

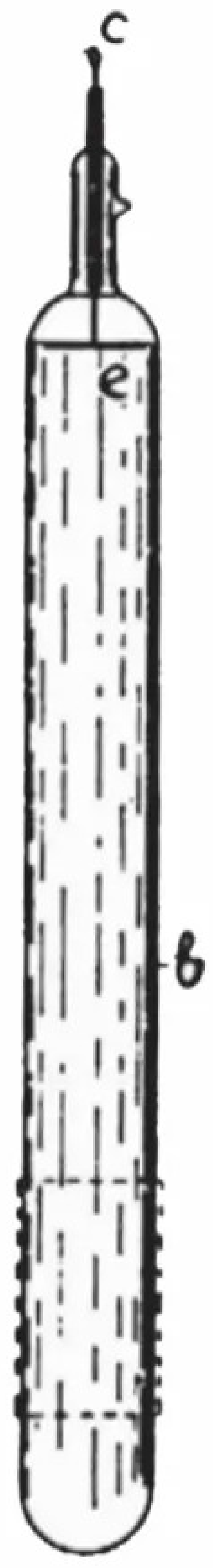


Fig.10: the "unipolar vacuum tube" comprising a glass bulb (b), a single electrode (e) and a lead-in conductor (c). A second electrode could be added towards the bottom; otherwise, the return circuit was via capacitive coupling through the air. *Source: Tesla Universe* – https://pemag.au/link/abyf

T1's secondary to the lower (90°) end of T2's secondary and adding a tap at $\sqrt{3} \div 2$ or 86.6% of T2's secondary. The third phase requires no extra connections to generate as the 240° waveform is simply available as 360° – 120° (360° = 0°), so between the start of T1's secondary and that same 120° tap.

This can also work in reverse, to convert three-phase to two-phase, but in that case the load has to be perfectly balanced, as it would be in a motor.

Wireless World System

1893 and later

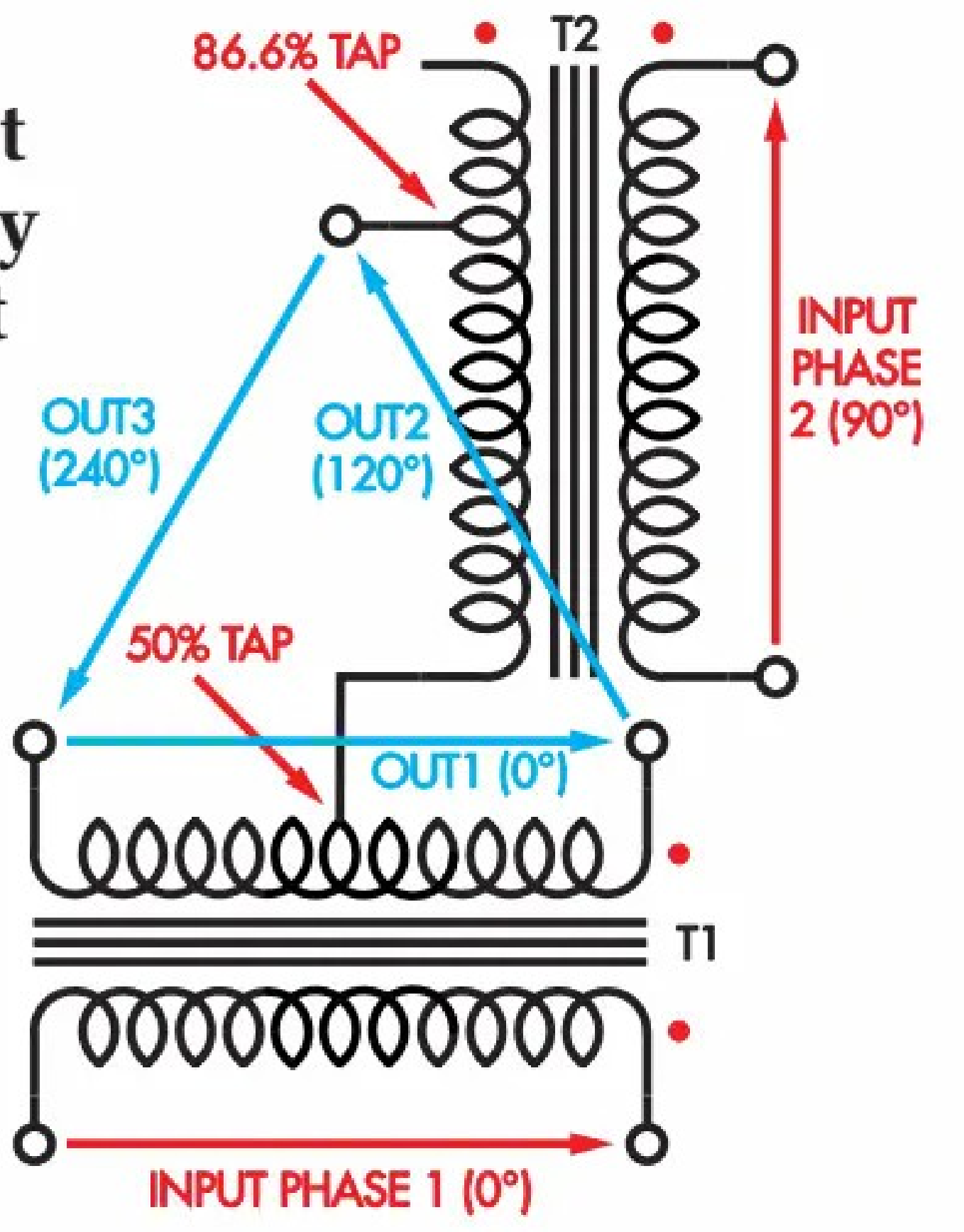
In 1893, he established the foundations of what he would call in a 1900 brochure, "The Wireless World System". It was to be a global wireless communications and wireless power transmission system (see Fig.13). According to Tesla, it would allow "the transmission of electric energy without wires" as well as point-to-point communications.

He said that the communications aspects of the system would allow "the instantaneous and precise wireless transmission of any kind of signals, messages or characters, to all parts of the world." and "... an inexpensive receiver, not bigger than a watch, will enable him [the user] to listen anywhere, on land or sea, to a speech delivered, or music played in some other place, however distant".

In 1915, in the New York Times, he added that the system "would enable thousands of persons to talk at once between wireless stations and make it possible for those talking to see one another by wireless, regardless of the distance separating them" (see page 136 of the PDF at https://pemag.au/link/aby0). All that sounds very familiar today!

To implement this system, he convinced banker John Pierpoint Morgan (J.P. Morgan) to invest in this project, which he built at Wardenclyffe on Long Island, New York (more on that next month).

Fig.12: a simple but clever way to convert two-phase AC to three-phase.


86.6% TAP
T2
INPUT PHASE 2 (90°)
OUT3 (240°)
OUT2 (120°)
50% TAP
OUT1 (0°)
T1
INPUT PHASE 1 (0°)

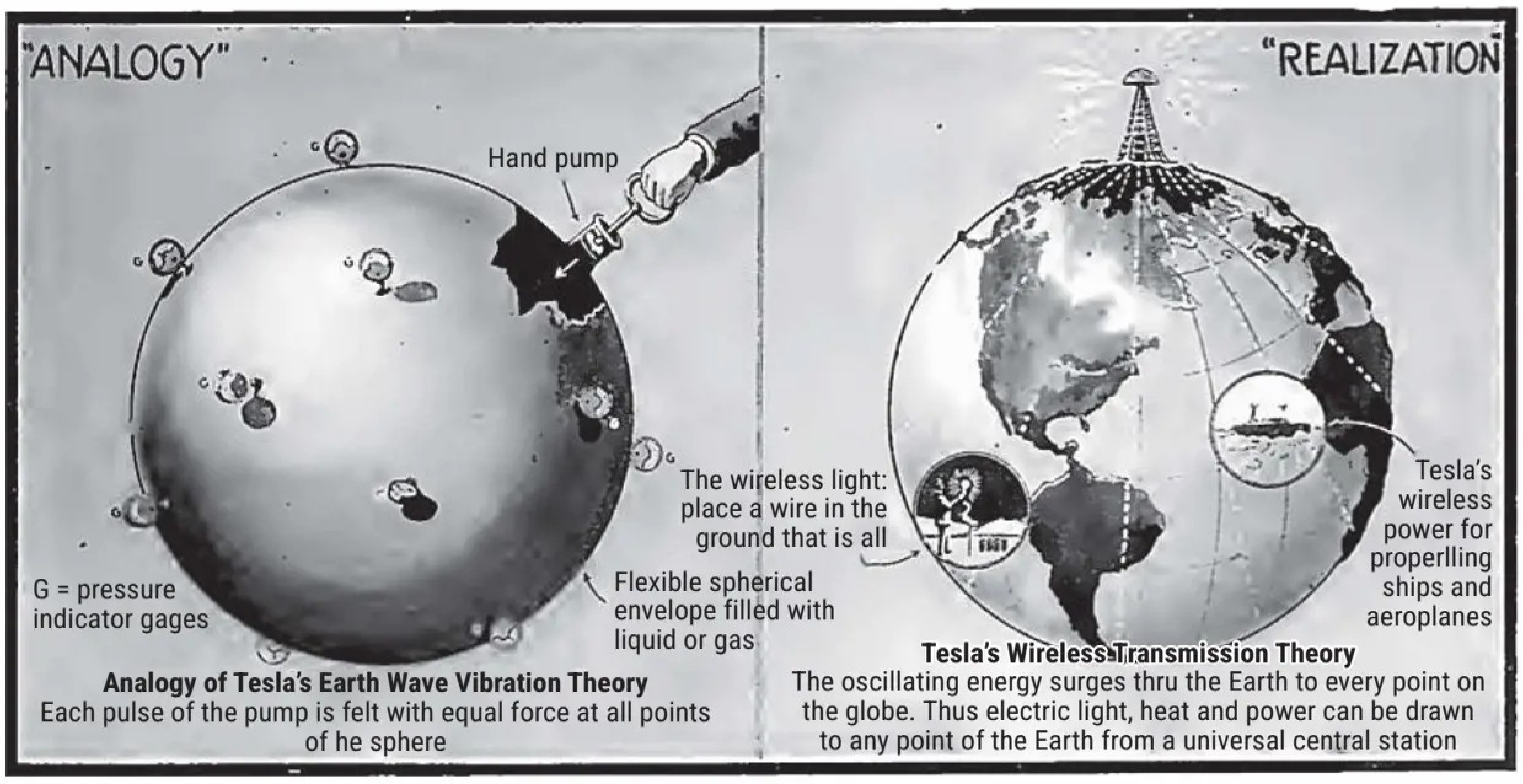

"ANALOGY"
Hand pump
G = pressure indicator gages
Flexible spherical envelope filled with liquid or gas
Analogy of Tesla's Earth Wave Vibration Theory
Each pulse of the pump is felt with equal force at all points of he sphere
"REALIZATION
The wireless light: place a wire in the ground that is all
Tesla's wireless power for properlling ships and aeroplanes
Tesla's Wireless Transmission Theory
The oscillating energy surges thru the Earth to every point on the globe. Thus electric light, heat and power can be drawn to any point of the Earth from a universal central station

Fig.13: Tesla's proposed scheme to deliver light and power anywhere on Earth by "ground waves" travelling through the Earth. Illustration by Tesla from *Electrical Experimenter*, February 1919. *Source:* https://w.wiki/AZN8

Tesla received numerous patents for wireless communications and power transmission, such as transformer design, transmission methods, tuning circuits and signalling methods. Tesla also envisaged a system of thirty telecommunications towers worldwide, linked to telegraph and telephone systems.

He proposed transmitting "radiations", which were not Hertzian waves and would apparently travel through the Earth with little loss. The energy of such waves could be harnessed anywhere on Earth simply by placing a wire in the ground. We now know that radio waves do not travel through the Earth to any significant degree.

A reciprocating engine

1894

In 1894, Tesla received US patent 514,169 for a multipurpose reciprocating engine device that used gas or

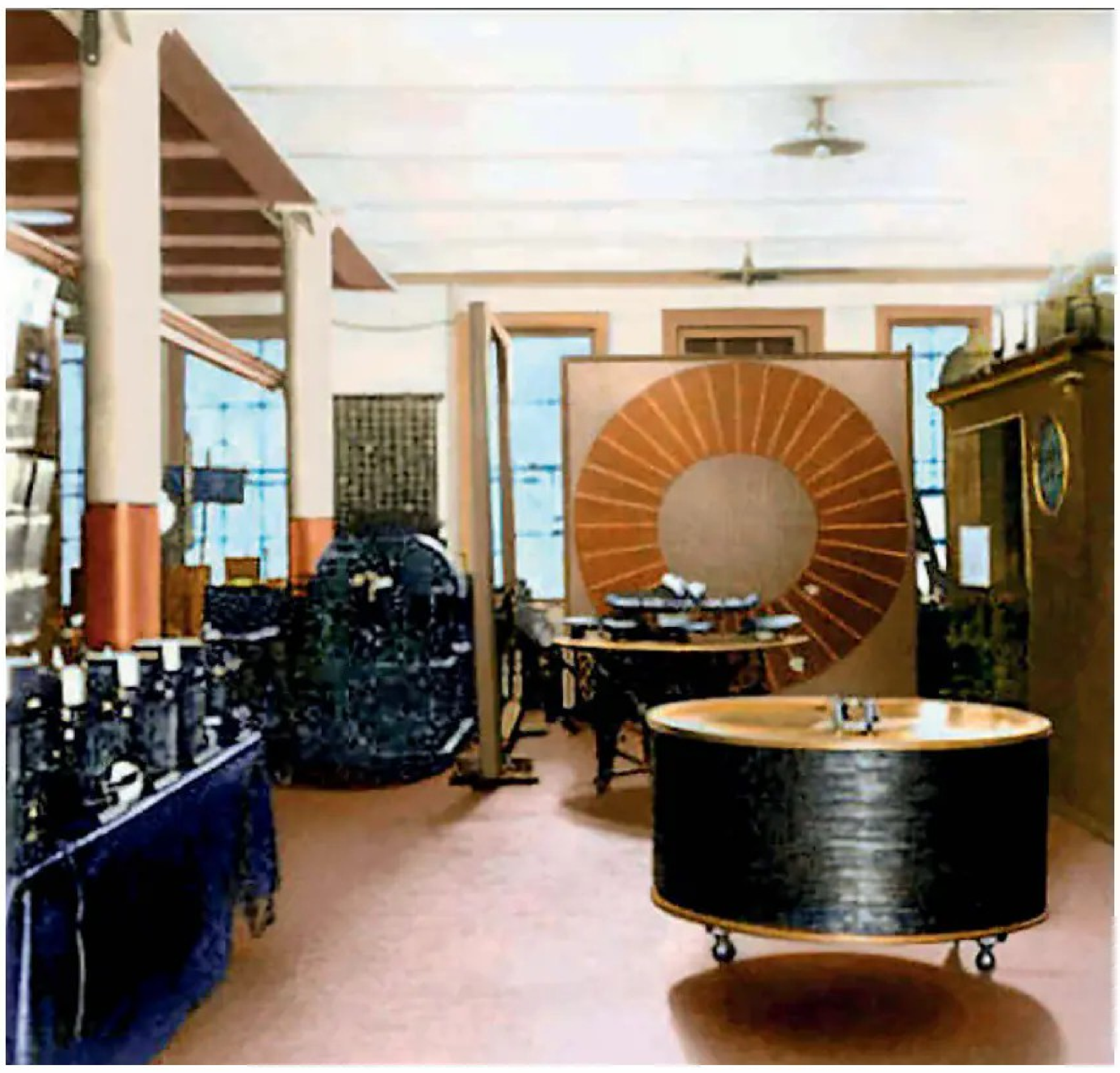


Fig.11: a colourised photo of the interior of Tesla's East Houston St lab. It is lit by lights of Tesla's design. *Source:* https://teslaresearch.jimdofree.com/labs-in-new-york-1889-1902/

steam pressure to generate mechanical oscillatory movement to generate electricity or for other purposes. The development of efficient steam turbines rendered its use for electricity generation obsolete.

In the New York World-Telegram of July 11th, 1935, Tesla recounted an incident of 1887 or 1888 where he said a little version of this device apparently brought an entire building to resonance, potentially destroying the building had he not stopped the it with a hammer (**www.rexresearch.com/teslamos/tmosc.htm**).

Mythbusters looked at this in their episode on "Nikola Tesla's Earthquake Machine!" (season 4, episode 20 – **https://youtu.be/LHsHiKtjoag**).

X-rays
1894

In 1894, Tesla worked with Crookes tubes and a "unipolar vacuum tube" of his own design. He noted mysterious damage to photographic plates in his laboratory by some sort of radiant energy.

Although X-rays had yet to be discovered or named, Tesla realised the source of the damage was rays from the point where the "cathodic stream" in the devices struck the anode (Crookes tube) or glass wall (his tube) – see Fig.10. It was later discovered that such a process generates X-rays.

In 1895, Wilhelm Röntgen discovered and published work about this "new kind of rays" (X-rays). Tesla started to work on X-rays and, in 1896, he reported being able to produce radiographs at a distance of ~12m; see **https://pemag.au/link/aby0** (p33).

Had Tesla fully recognised the phenomenon causing damage to his photographic plates, he may have been credited with their discovery. Tesla gave Röntgen full credit for his discovery. Later versions of Tesla's unipolar vacuum tube had a cooling system.

Laboratory fire
1895

In March 1895, Tesla's laboratory at South Fifth Ave, New York (occupied 1892-1895) burned to the ground. Tesla lamented, "I am in too much grief to talk. What can I say? The work of half my lifetime, very nearly all my mechanical instruments and scientific apparatus, that it has taken years to perfect, swept away in a fire that lasted only an hour or two... Everything is gone. I must begin over again."

This is said to have delayed his application for radio patents.

Wireless power experiments
1895

In his East Houston Street laboratory (1895-1902), he conducted experiments on the wireless transmission of electricity, setting up large Tesla coils, other types of resonant transformers and other apparatus (see Fig.11). He was producing up to 4MV, the maximum he could safely work with in a city building.

The Nikola Tesla Company
1895

In 1895, the Nikola Tesla Company was set up to fund, develop, and market Tesla's patents, which it did for the next few decades.

Transformers/induction coils
1897

In 1897, he was granted US patent **593,138** for a safe high-voltage, high-frequency electrical transformer/induction coil. In this patent, he showed single-wire electricity transmission with the return circuit flowing through the Earth. This concept was recently demonstrated in 2023, when 5kW was transmitted over 5km with 87% efficiency; see **https://ieeexplore.ieee.org/document/10023995**

Application for radio patents
1897

In 1897, Tesla applied for US patents **645,576** and **649,621**, both granted in 1900. These are considered his first radio patents, which he stated were relevant to "energy of many thousands of horsepower [being] transmitted over vast distances". In other words, he thought large amounts of electrical power could also be transmitted via this technology.

At this stage, wireless power transmission was his main focus, rather than radio communications.

Contrary to popular belief, Tesla did not invent radio and, unfortunately, did not have a good or correct understanding of the physics involved. In 1900, Guglielmo Marconi also applied for a US patent for radio, but it and subsequent revisions were rejected based on Tesla's preexisting patents.

However, in 1904, Marconi was granted US Patent **757,559** for radio, which he had applied for in 1901. Marconi had also previously applied for a British patent for radio in 1896, and it was granted in 1897, predating Tesla. The British patent (**12,039**) was the first for a system of wireless telegraphy using Hertzian waves.

Marconi was thus recognised as the inventor of radio; he shared a Nobel prize for it in 1909 with Karl Braun. Of course, there were many other contributors to the invention of radio, such as Reginald Fessenden, Heinrich Hertz, Oliver Lodge and John Stone.

There was litigation over early radio patents, and in 1943, the US Supreme Court settled a case involving Tesla's patents. However, it was not, as is often claimed, a case about who invented radio but who would be compensated by the US Government for using various patents during WW1. The story is too complicated to go into here; see **https://earlyradiohistory.us/tesla.htm**

Next month

In the following article, we'll pick up where we left off and cover the remainder of Tesla's life, from 1898 until his passing in 1943. We'll then go over related topics such as Tesla's mistakes and misconceptions, why the World Wireless System could never work, the 'war of the currents' and Tesla's lost files. PE

Practical Electronics PCB SERVICE

PROJECT	CODE	PRICE
OCTOBER 2025		
Seaside Sound Simulator PCB (blue)	01111241	£7.99
Fender Jazz Bass pickguard	23109241	£5.99
J&D Luthiers T-Style pickguard	23109242	£5.99
Music Man Stingray pickguard	23109243	£5.99
Fender Telecaster pickguard	23109244	£4.99
3D Printer Filament Dryer control PCB	28110241	£4.99
SEPTEMBER 2025		
Automatic LQ Meter PCB	CSE240203A	£5.99
Automatic LQ Meter front panel (black)	CSE240204A	£5.99
Compact Hi-Fi Headphone Amplifier PCB (blue)	01103241	£4.99
Discrete Ideal Bridge Rectifier TH PCB	18108241	£2.99
Discrete Ideal Bridge Rectifier SMD PCB	18108242	£2.99
Dual Rail Supply Protector PCB	18109241	£3.99
Audio Out Cascode Amplifier PCB	CASCODEA	£4.99
Audio Out Cascode Amplifier Power Supply PCB	CASCODEP	£4.99
AUGUST 2025		
Micromite Explore-40 PCB	07106241	£2.99
Micromite Explore-40 Audio Breakout PCB	07101222	£2.99
8-Channel Learning IR Remote Receiver PCB	15108241	£5.99
FlexiDice PCB set	08107241-2	£4.99
JULY 2025		
Compact OLED Clock & Timer PCB	19101231	£4.99
180-230V DC Motor Speed Controller PCB	11104241	£9.99
Mains Power-Up Sequencer PCB (same as Feb '25)	10108231	£11.99
JUNE 2025		
Styloclone (case version)	23106241	£6.99
Styloclone (standalone version)	23106242	£7.99
Dual Mini LED Dice (TH LEDs, white solder mask)	08103241W	£2.99
Dual Mini LED Dice (TH LEDs, black solder mask)	08103241B	£2.99
Dual Mini LED Dice (SMD LEDs, white solder mask)	08103242W	£2.99
Dual Mini LED Dice (SMD LEDs, black solder mask)	08103242B	£2.99
Dual Hybrid Supply regulator board	18107211	£5.99
Dual Hybrid Supply front panel control board	18107212	£2.99
Dual Hybrid Supply CPU board	01106193	£3.99
Dual Hybrid Supply LCD adaptor	01106196	£2.99
MAY 2025		
Skill Tester 9000	08101241	£9.99
USB-C Serial Adaptor (black solder mask)	24106241	£2.99
DC Supply Protector (SMD adjustable version)	08106241	£3.99
DC Supply Protector (TH adjustable version)	08106242	£3.99
DC Supply Protector (TH fixed version)	08106243	£3.99
APRIL 2025		
ESR Test Tweezers PCB set	04105241+	£5.99
Fan Speed Controller Mk2	10104241	£4.99
Calibrated Microphone (SMD version)	01108231	£3.99
Calibrated Microphone (TH version)	01108232	£3.99
MARCH 2025		
Pico Gamer	08104241	£9.99
Compact Frequency Divider	04112231	£4.99
FEBRUARY 2025		
Laser Communicator transmitter	16102241	£3.99
Laser Communicator receiver	16102242	£3.99
Model Railway Points Controller transmitter	09101241	£3.99
Model Railway Points Controller receiver	09101242	£2.99
Mains Power-Up Sequencer	10108231	£11.99

PROJECT	CODE	PRICE
JANUARY 2025		
USB to PS/2 Converter for VGA PicoMite	07111231	£3.99
ps2x2pico USB to PS/2 Converter	07111232	£3.99
Compact Microphone Preamplifier (case version)	01110231	£5.99
Compact Microphone Preamplifier (integrated version)	01110232	£5.99
DECEMBER 2024		
Raspberry Pi Clock Radio PCB pair	19101241/2	£12.99
VGA PicoMite (blue)	07107221	£5.99
Secure Remote Switch receiver	10109231	£4.99
Secure Remote Switch transmitter (module version)	10109232	£2.99
Secure Remote Switch transmitter (discrete version)	10109233	£2.99
NOVEMBER 2024		
Ideal Diode Bridge Rectifiers (28mm square)	18101241	£3.99
Ideal Diode Bridge Rectifiers (22mm square)	18101242	£3.99
Ideal Diode Bridge Rectifiers (4-pin SIL)	18101243	£3.99
Ideal Diode Bridge Rectifiers (tiny SOT-23)	18101244	£3.99
Ideal Diode Bridge Rectifiers (standalone SMD)	18101245	£4.99
Ideal Diode Bridge Rectifiers (standalone TH; 2oz copper)	18101246	£6.99
Multi-channel Volume Control (volume PCB)	01111221	£4.99
Multi-channel Volume Control (control PCB)	01111222	£4.99
Multi-channel Volume Control (OLED PCB, black)	01111223	£4.99
OCTOBER 2024		
TQFP-32 Programming Adaptor	24108231	£4.99
TQFP-44 Programming Adaptor	24108232	£4.99
TQFP-48 Programming Adaptor	24108233	£4.99
TQFP-64 Programming Adaptor	24108234	£4.99
2m VHF FM Test Signal Generator	06107231	£4.99
SEPTEMBER 2024		
PIC Programming Adaptor	24105231	£4.99
Salad Bowl Speaker crossover	01109231	£7.99
AUGUST 2024		
Arduino ESR Meter combined L/C-ESR	04106182	£9.99
Arduino ESR Meter separate ESR	04106183	£7.99
Watering System Controller	15110231	£11.99
JULY 2024		
Dynamic NFC Tag (22 × 31mm)	06101231	£5.95
Dynamic NFC Tag (86 × 54mm)	06101232	£7.95
Reciprocal Frequency Counter	CSE230101C	£8.95
Reciprocal Frequency Counter front panel (black mask)	CSE230102	£7.95
Pi Pico-based Thermal Camera	04105231	£6.95
JUNE 2024		
Loudspeaker Test Jig	04106231	£12.95
MAY 2024		
GPS-Disciplined Oscillator	04103231	£12.95
Dual RF Amplifier	CSE220602A	£5.95
Songbird	08103231	£7.95
APRIL 2024		
Basic RF Signal Generator	CSE221001	£8.99
Basic RF Signal Generator front panel (black solder mask)	CSE220902B	£9.99
Silicon Chirp Cricket	08101231	£7.99
Test Bench 'Swiss Army Knife'	04110221	£9.99
MARCH 2024		
Digital Volume Control Potentiometer SMD version	01101231	£8.95
Digital Volume Control Potentiometer through-hole version	01101232	£9.95
Advanced SMD Test Tweezers body	04106221	£6.95
Advanced SMD Test Tweezers legs (pair)	04106212	£7.95

PCBs for most recent *PE/EPE* constructional projects are available. From the July 2013 issue onwards, PCBs with eight-digit codes have silk screen overlays and, where applicable, are double-sided, have plated-through holes, and solder mask. They are similar to photos in the project articles. Earlier PCBs are likely to be more basic and may not include silk screen overlay, be single-sided, lack plated-through holes and solder mask.

Always check price and availability in the latest issue or online. A large number of older boards are listed for ordering on our website.

In most cases we do not supply kits or components for our projects. For older projects it is important to check the availability of all components before purchasing PCBs.

Magazine back issues are available – see https://www.electronpublishing.com/product-category/pe-back-issues-print-and-digital/

Double-sided | plated-through holes | solder mask

PROJECT	CODE	PRICE
FEBRUARY 2024		
Active Mains Soft Starter	10110221	£10.95
JANUARY 2024		
Q Meter	CSE220701	£8.95
Q Meter front panel (black solder mask)	CSE220704	£7.95
Raspberry Pi Pico W BackPack	07101221	£8.95
DECEMBER 2023		
Digital Boost Regulator	24110224	£8.95
Dual-Channel Power Supply for Breadboards	04112221	£11.95
Display Adaptor for the Breadboard PSU	04112222	£9.95
NOVEMBER 2023		
LC Meter Mk3	CSE220503C	£9.95
DC Supply Filter for vehicles	08108221	£8.95
Discrete Audio Op Amp PCB	AO1-JUL23	£9.95
Discrete Audio Op Amp PCB + essential components	AO2-JUL23	£17.95
OCTOBER 2023		
Buck/Boost Charger Adaptor	14108221	£9.95
PIC Breakout Board for SOIC parts	24110225	£5.95
PIC Breakout Board for DIP parts	24110222	£5.95
AVR64DD32 Breakout board	24110223	£5.95
Automatic Train Controller	09109221	£6.95
Chuff Sound module	09109222	£6.95
SEPTEMBER 2023		
Mini LED Driver	16106221	£7.95
New GPS-Synchronised Clock	19109221	£9.95
AUGUST 2023		
110dB RF Attenuator	CSE211003	£9.95
Universal Battery Charge Controller (2023 update)	14107192	£9.95
Wide-Range OhmMeter	04109221	£12.95
JULY 2023		
Multimeter Checker/Calibrator	04107221	£10.95
JUNE 2023		
Arduino Programmable Load	04105221	£8.95
Buck-Boost LED Driver	16103221	£8.95
MAY 2023		
Precision AM-FM DDS Signal Generator	CSE211002	£14.95
Improved SMD Test Tweezers programmed PIC	0410621PIC	£12.95
AO 2x-dual-to-quad through-hole (pack of five)	AO1-MAY23	£6.95
AO 2x-dual-to-quad SMD (pack of five)	AO2-MAY23	£6.95
AO 2x-single-to-dual through-hole (pack of five)	AO3-MAY23	£6.95
AO 2x-single-to-dual SMD (pack of five)	AO4-MAY23	£6.95
APRIL 2023		
500W Amplifier Module	see p.22, April 2023	
Clipping Indicator (per channel)	01112211	£7.95
CD Welder Power Supply (one needed)	29103221	£9.95
CD Welder Controller (one needed)	29103222	£9.95
CD Welder Energy Storage module (several needed)	29103223	£7.95
AO Universal Dual Op Amp Board	AO1-APR23	£9.95
AO Stereo RIAA precision passives kit for dual op amp	AO2-APR23	£8.95
MARCH 2023		
Pico BackPack	07101221	£9.95
Semaphore Signal (controller)	09103221	£7.95
Semaphore Signal (blade)	09103222	£5.95

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FEBRUARY 2023		
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Tesla Coil potentiometer board	26102222	£5.95
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Remote Gate Controller	11009121	£12.95
JANUARY 2023		
Classic LED Metronome – 8-LED	23111211	£7.95
Classic LED Metronome – 10-LED	23111212	£8.95
Multi-Channel Speaker Protector – 6 channel	01101221	£9.95
Multi-Channel Speaker Protector – 4 channel	01101222	£7.95
Remote Control Range Extender – IR-to-UHF	15109212	£5.95
Remote Control Range Extender – UHF-to-IR	15109211	£7.95
AO Universal Single Op Amp Board	AO1-JAN23	£7.95
DECEMBER 2022		
Hummingbird Amplifier	01111211	£9.95
SMD Trainer PCB	29106211A	£8.95
NOVEMBER 2022		
USB Cable tester – main PCB	04108211	£12.95
USB Cable tester – front panel	04108212	£5.95
USB Cable tester – optional panel	SC5970	£5.95
Model Railway Carriage Lights – PCB	09109211	£6.95
AO trnsfmr PCB – standard VTX-A range	VTX-101-007	£6.95
AO trnsfmr PCB – dual-outline VTX102-3001/101-3002	VTX-Dual	£6.95

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NOTE: Most boards are in stock and sent within seven days of order receipt. Please allow up to 28 days delivery if we need to restock.

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SMD Test Tweezers – PCB and pair of tweezer arms	04106211-2	£11.95
SMD Test Tweezers – programmed PIC12F1572-I/SN	0410621A	£7.95
Tele-com	12110211	£12.95
SEPTEMBER 2022		
Touchscreen Digital Preamp – main board	01103191	£12.95
Touchscreen Digital Preamp – adaptor board pair	01103192	£5.95
20A DC Motor Speed Controller	11006211	£9.95
AUGUST 2022		
Multi-purpose Battery Manager – I/O Expander	11104212	£5.95
Multi-purpose Battery Manager – Switch Module	11104211	£8.95
Simple MIDI Music Keyboard (for 8 switches)	23101213	£6.95
Nano Pong	08105212	£7.95
JULY 2022		
Silicon Labs AM/FM/SW Radio	CSE210301C	£10.95
Level Crossing Controller	09108211	£6.95
JUNE 2022		
Full-wave Motor Speed Controller	1010221	£8.95
PIC Programming Helper for 8-pin PICs only	24106211	£7.95
PIC Programming Helper for 8, 14 or 20-pin PICs	24106212	£10.95
Advanced GPS Computer	05102211	£9.95
MAY 2022		
Bus board PCB for Analogue Vocoder	AO1-MAY22	£10.95
Complete set of 14 PCBs for Analogue Vocoder	AO2-MAY22	£97.95
Programmed EEPROM for Digital FX Unit	FX1-MAY22	£10.95
Programmed PIC for Digital FX Unit w/ potentiometer	FX2-MAY22	£8.95
APRIL 2022		
64-key MIDI Matrix shield	23101211	£8.95
64-key MIDI Matrix switch board	23101212	£11.95
High-current Battery Balancer	14102211	£10.95
Digital FX Unit – using potentiometer	01102211	£9.95
Digital FX Unit – using BCD switch	01102212	£9.95
Universal Audio PSU	AO1-APR22	£11.95
MARCH 2022		
Mini Isolated Serial Link	24102211	£5.95
Busy Loo Indicator	16112201	£5.95
Analogue Vocoder – Band-pass filter board	AO1-MAR22	£9.95
Analogue Vocoder – HP/LP filter board	AO2-MAR22	£9.95
FEBRUARY 2022		
Arduino-based Power Supply	18106201	£9.95
Battery Monitor Logger	11106201	£10.95
Electronic Wind Chimes	23011201	£10.95
Analogue Vocoder – Driver Amplifier	AO-FEB22	£8.95
JANUARY 2022		
Vintage battery Radio Li-ion Power Supply	11111201	£9.95
MiniHeart: A Miniature Heartbeat Simulator	01109201	£8.95
DECEMBER 2021		
AM/FM/SW Digital Receiver	CSE200902A	£13.95
Balanced Input and Attenuator for USB CODEC	01106202	£11.95
NOVEMBER 2021		
Dual Battery Lifesaver	11111202	£6.95
OCTOBER 2021		
Mini Wi-Fi LCD BackPack	24106201	£8.95

PROJECT	CODE	PRICE
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USB SuperCodec PCB	01106201	£14.95
Audio DDS Oscillator PCB	01110201	£5.95
Audio DDS Oscillator rotary encoder	01110201-ENC	6.95
Programming Adaptor Board for Audio DDS Oscillator	01110202	£5.95
High-power Ultrasonic Cleaner main PCB	04105201	£14.95
High-power Ultrasonic Cleaner front-panel PCB	04105202	
Night Keeper Lighthouse PCB	08110201	£5.95
AUGUST 2021		
Ol' Timer PCB	19104201	£11.95
Ol' Timer 8x8 RGB LED module using WS2812B	19104201-88	£8.95
Ol' Timer set of acrylic case pieces and spacer	19104201-ACR	£8.75
Ol' Timer DS3231 RTC IC wide SOIC-16	19104201-RTC	£5.95
Wideband Digital RF Power Meter	04106201	£9.75
Switchmode 78xx regulators ***(PACK of 5!)***	18105201	£7.95
Cool Beans SMAD display	CB-AUG21	£11.95
JULY 2021		
ATtiny816 Breakout / Dev Board w/ Capacitive Touch	24110181	£9.75
IR Remote Control Assistant (Jaycar version)	15005201	£8.95
IR Remote Control Assistant (Altronics version)	15005202	£8.95
PIC18F Development Board	PNM-JUL21	£12.95
Microphone Preamplifier	AO-JUL21	£11.95
JUNE 2021		
Roadies' Test Signal Generator (surface-mount ver.)	01005201	£8.95
Roadies' Test Signal Generator (through-hole version)	01005202	£9.95
Touchscreen Wide-range RCL Box (Resistor module)	04104201	£18.95
Touchscreen Wide-range RCL Box (Ind/Cap module)	04104202	
KickStart Part 3 – Gyrator-based Audio Filter	KS3-2021	£7.95
MAY 2021		
7-Band Equaliser (Mono)	01104201	£8.95
7-Band Equaliser (Stereo)	01104202	£10.95
Car Altimeter	05105201	£7.95
APRIL 2021		
Reflow Oven – DSP Active Crossover (CPU)	01106193	£19.95
Reflow Oven – DSP Active Crossover (Front panel)	01106195	
Reflow Oven – DSP Active Crossover (LCD)	01106196	
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MARCH 2021		
Nutube Guitar Effects Pedal	01102201	£12.95
Programmable Thermal Regulator (Peltier Interface)	21109181	£18.95
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Tunable HF Preamp	CSE190502	£8.95
FEBRUARY 2021		
4G Remote Monitoring	27111191	£9.95
JANUARY 2021		
Nutube Valve Preamplifier	01112191	£12.95
Arduino DCC Controller	09207181	£10.95
DECEMBER 2020		
Pseudo-Random Sequence Generator	16106191	£7.95
Clever Charger	14107191	£11.95
PE Theremin Amplifier	AO-1220-01	£8.95
NOVEMBER 2020		
LED Christmas Tree (1 off)	16107181-1	£6.95
USB/SPI Interface Board	16107182	£5.95
45V/8A Power Supply PCB *plus* acrylic spacer	18111181	£14.95
45V/8A Power Supply five-way display bezel	18111181-BZ	£3.95
Five-way LCD Panel Meter/Display	18111182	£7.95

Silicon Chip kits & parts for projects in this issue (posted from Australia):

Compact Hi-Fi Headphone Amplifier kit (including case)	https://siliconchip.au/Shop/20/6885	AUD $84 (about £42.00) + P&P

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ADVERTISING INDEX

AOSHOP 13
BACK ISSUE PDFs. 33
BASIC MICRO 26
ENIGMA COMPONENTS SHOP 19
HAMMOND MANUFACTURING. 40
MICROCHIP TECHNOLOGY . . . . Outer back cover
PCBWAY. . . . . Inner front cover
TEACH-IN BUNDLES 59
WIRELESS FOR THE WARRIOR BOOKS 19, 45

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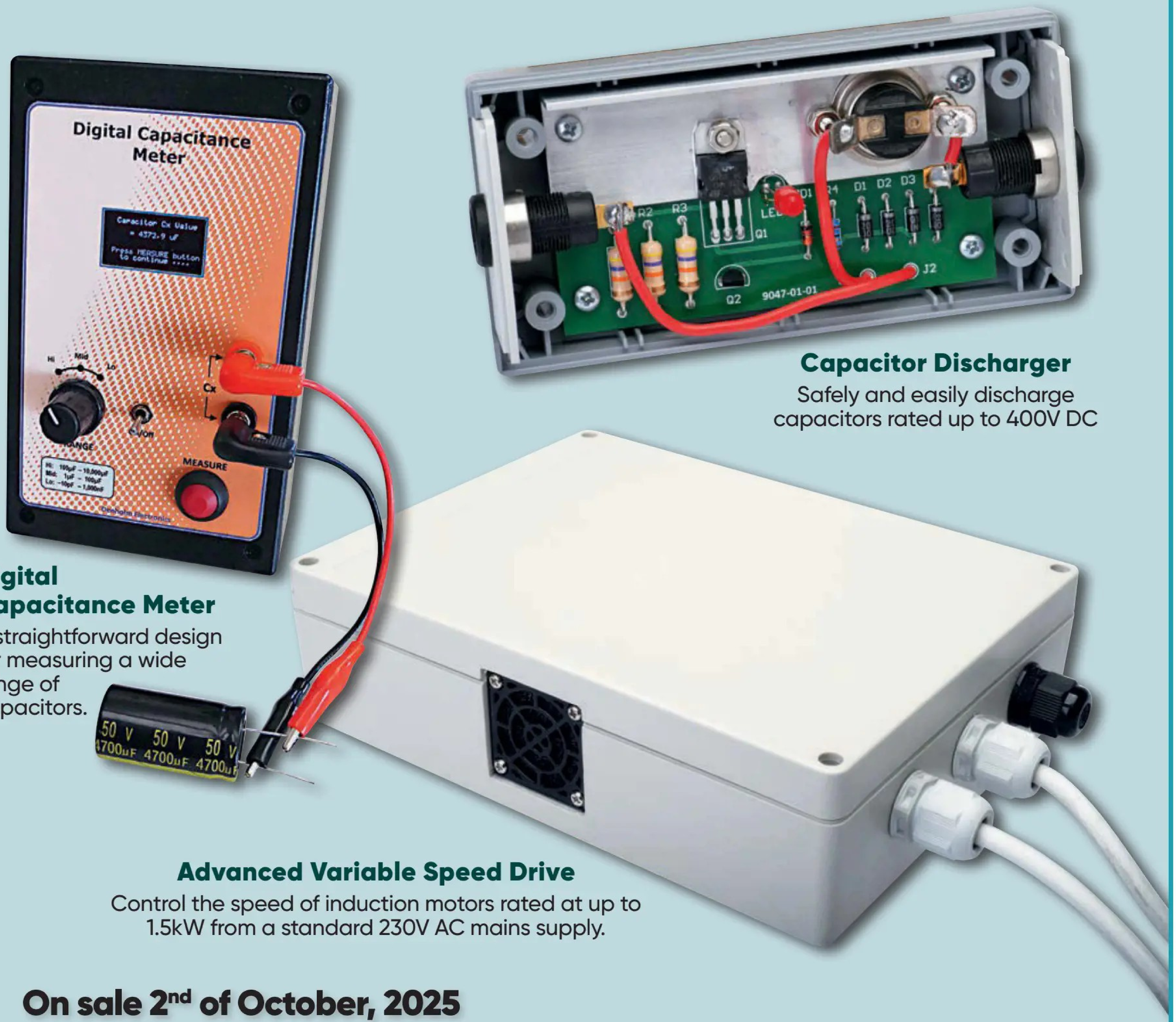
Next Month – in the November issue



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